

Measuring insertion loss of cavities

By Harold Kinley, C.E.T.

How much insertion loss is your cavity causing at the *desired* or *pass* frequency? How does a field technician measure the insertion loss without access to laboratory-type equipment or under field-test conditions? It can be done with reasonable accuracy using some basic methods.

Recently, while surfing the Internet, I came across an interesting application note in the form of a "Tech-aid" called "Measuring Insertion Loss with Wattmeters" from TX RX Systems, Angola, NY. You can find it at www.txrx.com/f3c.html. With permission from TX RX Systems, the document from the Web site is presented here, with minor editing for space considerations. Following the note, I offer some of my thoughts. (On the Web site, Figure 1 and Figure 2 are the same, which is an obvious oversight. So, I added Figure 1 to match the text description of the test setup.)

MEASURING INSERTION LOSS WITH WATTMETERS

Passband insertion loss is a primary specification of a wide variety of passive RF devices. Typical insertion loss values are relatively small and therefore are difficult to measure with anything

but laboratory-quality instruments. At TX RX Systems, we measure insertion loss using the latest Hewlett-Packard network analyzers. The calibration of our instruments is traceable to the National Institute of Standards and Technology (NIST) ...

It is not easy to accurately measure low values of insertion loss at high power using ordinary radio transmitters as signal sources and RF wattmeters, such as the Bird model 43 or similar. In fact, insertion loss under power cannot be measured directly; it must be calculated as a ratio of measured RF power at the output and input of the device under test (DUT), using the formula:

$$IL = 10 \times \log (P_o \div P_i)$$

where IL is insertion loss (in decibels), P_i is RF power input at the DUT and P_o is RF power at output of the DUT.

This Tech-aid explains why indirect insertion loss measurements made with a single wattmeter are prone to significant errors and describes a two-wattmeter method that is more likely to be in agreement with measurements made with laboratory-quality instruments.

Single-wattmeter method

In general, it is not possible to use a single wattmeter to measure device

input and output power with sufficient accuracy to verify factory insertion-loss specifications. There are just too many possible sources of error.

Figures 1A and 1B describe a common method of performing the required power measurements. The DUT has a factory-measured insertion loss of -1.5dB at the pass frequency. The wattmeter is a Bird model 43 with a 50W element, and the transmitter is a 30W mobile transceiver. Random-length coaxial cables are used for equipment connections.

In Figure 1A (below left), the transmitter is connected to the DUT via the wattmeter and test cables 1 and 2. The output of the device is connected to a 50Ω load via cable 3. When the transmitter is keyed, the wattmeter indicates 32.3W forward power. We record $P_i = 32.3\text{W}$. In Figure 1B (below left), the transmitter is connected to the DUT via cable 1. The output of the DUT is now connected to the load via cable 3, the wattmeter and cable 2. The wattmeter now indicates 20W forward power. We record $P_o = 20.0\text{W}$.

With the above power measurements, insertion loss is calculated as follows:

$$\begin{aligned} IL &= 10 \times \log (20 \div 32.3) \\ &= 10 \times (-0.21) \\ &= -2.1\text{dB} \end{aligned}$$

Is this filter out of specification? Before we reach any conclusions, let us take a close look at the possible sources of error inherent to the single-wattmeter method.

Sources of measurement error

► Transmitter load impedance changes — Substantially different lengths of cable are used to connect the DUT to the transmitter in 1A and 1B. If filter input impedance is not purely resistive and equal to 50Ω, changing the length of cable between it and the transmitter can change the magnitude and phase of the load impedance presented to the transmitter. Transmitter output power may therefore change, due to the

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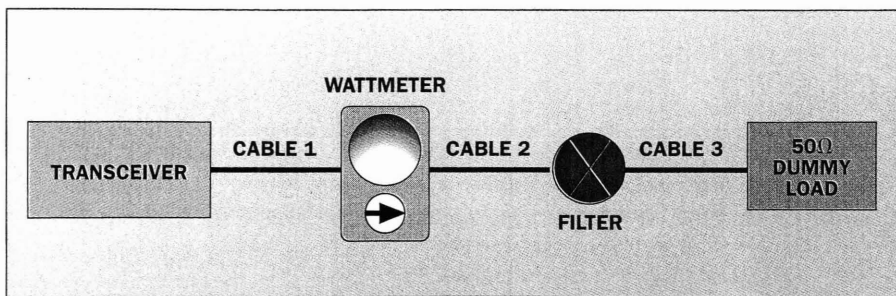


Figure 1A. The wattmeter is positioned on the input side of the filter and the forward power is measured. Random lengths of cable are used.

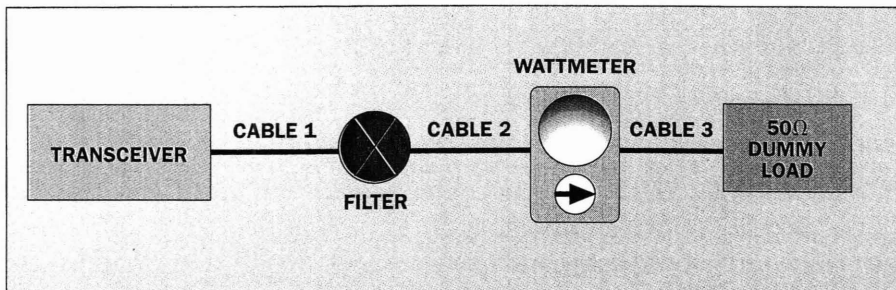


Figure 1B. The forward power on the output side of the filter is measured in this setup.

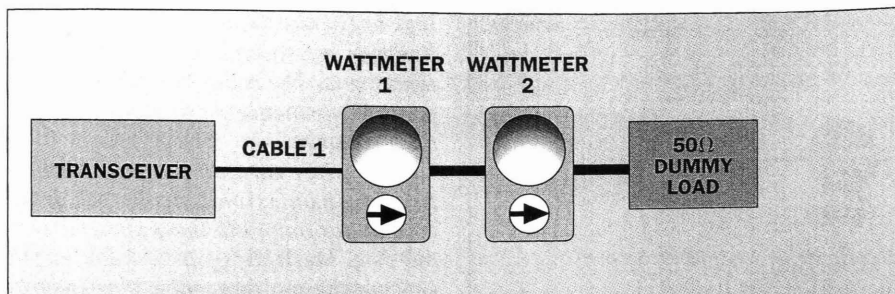


Figure 2A. This setup is used to establish the calibration of the wattmeters and to determine a correction factor.

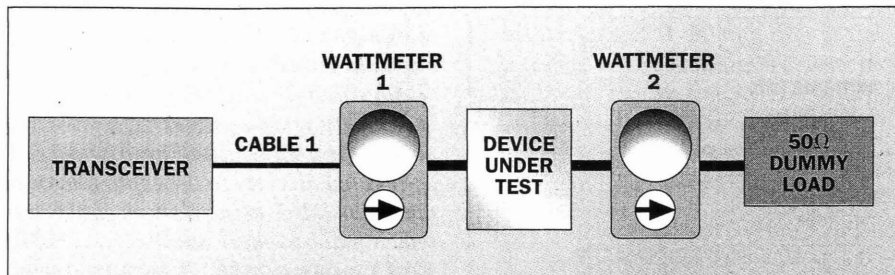


Figure 2B. This setup is used to determine the insertion loss of the filter or other DUT (device under test).

load impedance changes caused by moving the wattmeter and one cable from the input to the output of the DUT.

► **Wattmeter position** — Standing waves exist in a transmission line that is terminated in a mismatched or reactive load. Depending on the VSWR, different

power measurements may be obtained with the wattmeter at different points within the standing wave pattern in the test cables.

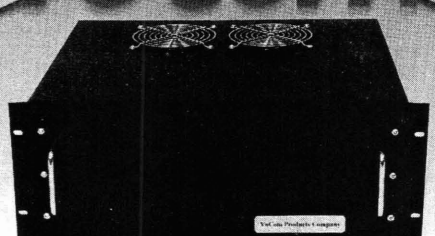
► **Cable and fitting insertion loss** — The loss of all interconnecting cables and fittings that affect power measurements

must be considered in the calculation of insertion loss. The total error ... could easily be about 0.6dB, the discrepancy between factory specification and field measurement ... worse if the test transmitter were unstable.

► **Transmitter instability** — Some transmitter power amplifiers may not be stable at all load impedances and phase angles. Resonant devices in particular may exhibit large reactive components at off-resonance frequencies. This may induce parametric oscillations that produce significant output power at frequencies outside the passband of the DUT. If the transmitter is oscillating, a wattmeter measurement of transmitter output power will include spurious power. If spurious power is significantly attenuated by the filter under test, a "false insertion loss" will result. Depending on the ratio of spurious to carrier power, and the response of the DUT, even larger insertion loss measurement errors may be induced.

In Figure 1C [not shown], the transmitter oscillates when it is connected to the DUT via a shorter, unfavorable length of cable. The wattmeter on the filter output reads only 15.5W forward power, because about 4.5W of spurious power are not passing through the cavity

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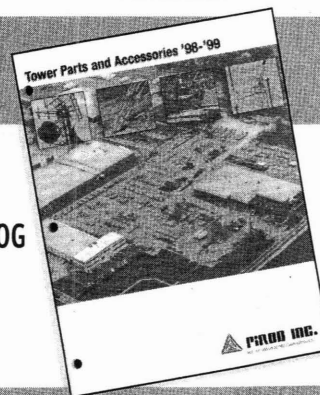
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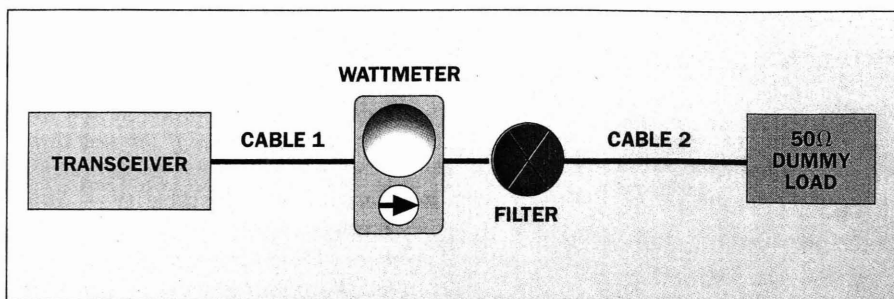


Figure 3A: This single-wattmeter method of measuring insertion loss is valid as long as proper attention is paid to cable lengths. The input power to the filter is measured here.

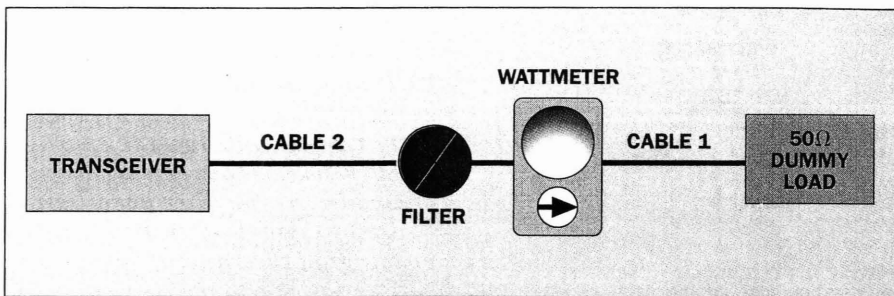


Figure 3B: The output power from the filter is measured here. Cable lengths of one-half wavelengths (or multiples) are maintained in this procedure.

filter. Insertion loss now appears to be:

$$\begin{aligned} IL &= 10 \times \log (15.5 \div 32.5) \\ &= 10 \times (-0.32) \\ &= -3.2\text{dB}. \end{aligned}$$

This result is completely wrong.

Recommended test method

TX RX Systems recommends a two-wattmeter, two-step method that eliminates most of these problems. The arrangement in Figure 2A (see page 49) is first used to obtain wattmeter read-

ings to correct for test equipment insertion loss and wattmeter relative calibration errors. The setup in Figure 2B (see page 49) is then used to measure input and output power.

Cable 1 is cut to a length such that total transmission line length from transmitter output to the wattmeter output is a multiple of a $\frac{1}{2}$ -wavelength ($\lambda/2$) at the test frequency. This ensures that the load impedance "seen" by the transmitter is essentially the same as the impedance of the DUT connected to wattmeter 1. Wattmeter operating manuals usually contain information on optimal lengths of cable required for various frequency ranges.

In Figure 2B, wattmeters 1 and 2 are connected directly to the input and output of the DUT, using a short cable or a type N male-to-male union (UG-57B/U, TX RX part no. 8-5857). It may also be necessary to use a type N, 90° male-to-female adapter (UG-27C/U, TX RX part no. 8-5867) to facilitate installation of the wattmeters near the DUT. In Figure 2A, the wattmeters are connected using the same fittings as in Figure 2B, plus an additional type N female-to-female union (UG-29B/U, TX RX part no. 8-5856).

The 50Ω load resistor in Figures 1 and 2 should be connected to the output of wattmeter 2 via a short length of coaxial cable or a type N male-to-male union. The length of transmission line between the wattmeter and the load resistor is not critical if return loss is -30dB or better.

Test procedure

1. Connect wattmeter 1 directly to wattmeter 2 as shown in Figure 2A. Key the transmitter and record the forward power readings, P1 and P2.
2. Unkey the transmitter and insert the DUT between wattmeters 1 and 2, as shown in Figure 2B.
3. Key the transmitter again and record the forward power readings, P3 and P4.
4. Calculate insertion loss as follows:

$$IL = 10 \log ((P1 \times P4) \div (P2 \times P3)).$$

The following measurements are obtained with the cavity filter in the previous example: P1 = 28.7W; P3 = 28.0W; P2 = 24.0W; and P4 = 16.8W.

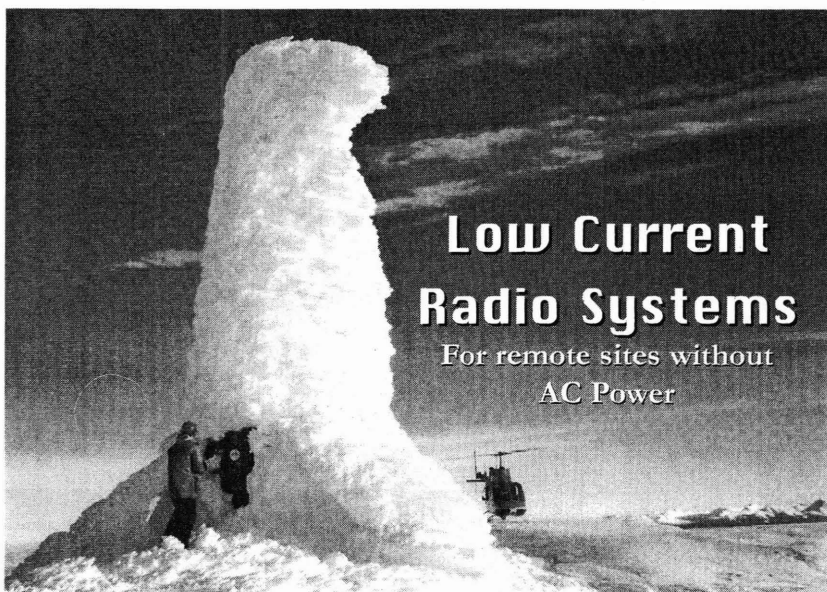
Computed insertion loss is:

$$Po \div Pi = (28.7 \times 16.8) \div (24.0 \times 28.0) = 0.7175$$

$$\begin{aligned} IL &= 10 \log (0.7175) \\ &= -1.44\text{dB} \end{aligned}$$

This is within less than 0.1dB of the factory specification for the filter under test.

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In case of a discrepancy, if you use the suggested two-wattmeter method and your insertion loss measurement does not agree with ours, there are two likely causes: Either your transmitter is oscillating, or our equipment is indeed out of specification.

With the equipment arranged as in Figure 2B, insert a -30dB RF sampler on the transmitter output and check the transmitter spectral purity with a spectrum analyzer. Carefully scan a broad frequency range, as spurious outputs may sometimes appear at frequencies far removed from the carrier. If the transmitter is clean, it is safe to conclude that the equipment is out of specification. You should then file a warranty claim.

Other precautions

1. Use wattmeter elements that produce a wattmeter reading at mid-scale or higher.
2. Use the best cable and connectors money can buy. If possible, connectors should be crimped and soldered. A bad connector will ruin your measurement.
3. Do not use UHF connectors or adapters. UHF connectors are notorious for their bad impedance characteristics even at VHF frequencies. Put type N connectors on your wattmeter.
4. Check the spectral purity of your test transmitter. All spurious products and harmonics must be at least 60dB to 70dB below the carrier when the transmitter is loaded as in Figure 2B.

Brief observations

The points made by this application note are well worth remembering and following. I would like to comment on a couple of points. Concerning the wattmeter position along the line, the note stated that "Depending upon the VSWR, different power measurements may be obtained with the wattmeter at different points within the standing wave pattern in the test cables." Specific mention was made of the Bird model 43 wattmeter. The Bird 43 instruction book, page 2-0, states "Being highly directional, the ThruLine Element is sensitive at one setting only to one of the traveling waves, which produce standing waves by interference. ThruLine measurements are therefore independent of position along standing waves. It may be said that the ThruLine doesn't know, doesn't care and doesn't need to care where it is along a standing wave."

Because any spurious signal would affect the single-wattmeter method of

measuring insertion loss, it would also affect the two-wattmeter method. The two-wattmeter method gives no advantage in this regard.

Single-wattmeter method

I am not arguing against the two-wattmeter method of performing the insertion loss test. However, if only one wattmeter is available, there is a way to use a single wattmeter to measure the insertion loss of a filter with a reasonable degree of accuracy. Refer to Figures 3A and 3B on page 50.

In Figure 3A (assuming N connectors throughout), the wattmeter is connected to the filter input directly using an N male-to-male connector. The length of cable 1 is chosen such that:

cable 1 + wattmeter through-line section + N male-to-male connector

is equal to $\lambda/2$ (or a multiple thereof) at the operating frequency. The type N male-to-male connector is used to connect the wattmeter directly to the filter in both Figures 3A and 3B. The velocity factor of the cable must be taken into account when calculating the cable length. Assuming that all the lines between the transmitter output and the filter input total $\lambda/2$, or a multiple thereof, the input impedance of the filter will be seen by the transmitter output.

Cable 2, connecting the filter output to the dummy load, is also cut to $\lambda/2$ (or a multiple).

With the setup in Figure 3A, the transmitter is keyed and the forward power reading is recorded. Record this measurement as P_i . Also take a look at the reverse power reading to check for a severe impedance mismatch. Now, using the setup shown in Figure 3B, key the transmitter and record the wattmeter reading as P_o . Notice that in Figure 3B, cable 2 connects the transmitter to the filter input. The line length is still $\lambda/2$, so the impedance seen by the transmitter has not changed.

Using the readings, you can determine the insertion loss by the formula:

$$IL = 10 \log (P_i \div P_o).$$

If $P_i = 40W$ and $P_o = 30W$, then the insertion loss of the filter is:

$$10 \log (40 \div 30) = 10 \log (1.333) = 1.25dB.$$

All precautions stated for the two-wattmeter method apply to the one-wattmeter method as well. Reasonable care in setting up the test procedure will yield reasonable accuracy in the result.

Until next time—stay tuned!

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Danger: High voltage

Voltage mismatches can be harmful

When a transmission line is terminated by an impedance other than a purely resistive imped-

ance equal to the characteristic impedance of the line, standing waves exist on the line. When a high voltage standing wave ratio (VSWR) exists on a transmission line there are points of voltage maxima (antinodes) and voltage minima (nodes) — the greater the mismatch or VSWR, the greater the peak voltage on the line. If the voltage at the peaks or maxima becomes high enough, punch-through can occur. Punch-through is the point at which the voltage between the inner conductor and the shield (outer conductor) exceeds the voltage rating of the dielectric material or insulation of the center conductor.

It is important to know the characteristics of the transmission line and to be able to calculate the peak voltage that can be expected on the line for a given VSWR or load mismatch condition. Furthermore, if the reactive and resistive parts of the load impedance are known, it is possible to calculate the exact location on the line where the peak voltage will occur. It is important to remember that any reactive load will produce standing waves on the line.

Quantifying the mismatch

Many units of measure are used to define or quantify the degree of mismatch on a line. Reflection coefficient, SWR, return loss, and reflected power (in percentage) are used to quantify the degree of mismatch on the transmission line. Typically, the land mobile radio technician uses a directional wattmeter connected in the transmission line to measure the for-



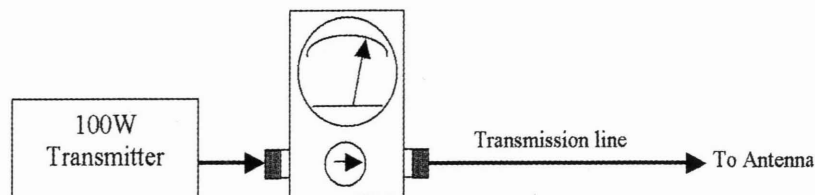
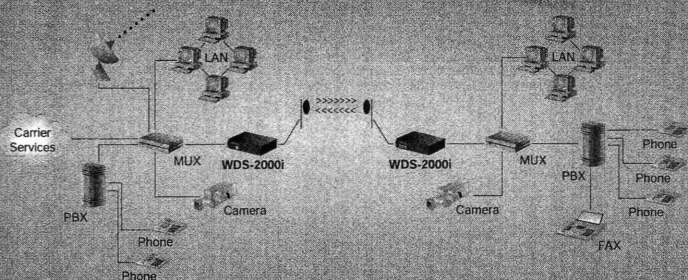


Figure 1.

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ward and reflected power. If the load impedance perfectly matches the transmission line impedance, the reflected power is zero. If the load impedance is not equal to the transmission line impedance, the wattmeter will indicate the amount of reflected power. The net power is equal to the difference between the forward power and the reflected power. The net power is the amount of power that is dissipated or radiated — usually a combination of both, since line losses are incurred in any practical transmission line.

Suppose that the setup shown in Figure 1 indicates a forward power of 100W and a reflected power of 25W. Assume for the moment that the transmission line loss is zero. In this case, the forward power at the load (antenna) is 100W and the reflected power at the antenna is 25W. In terms of percentage, the reflected power is 25 percent at the antenna and the transmitter. The VSWR is the same at all points on the line.

Reflection coefficient can be anything from zero to one, with zero representing a perfect match and one representing the worst-case mismatch. If the forward and reflected power is known the reflection coefficient can be found from the formula below, where Γ (gamma) represents the reflection coefficient and r is the reflected-to-forward power ratio.

$$\Gamma = \sqrt{r}$$

In the case above where the forward power was 100W and the reflected power was 25W, the reflected-to-forward power ratio, r , is 0.25. Plugging this into the formula above yields:

$$\Gamma = \sqrt{0.25} = 0.5$$

The reflection coefficient can also be calculated from the SWR. The formula is shown below, where S represents SWR and G is reflection coefficient.

$$\Gamma = \frac{S-1}{S+1}$$

If the SWR is 3, plugging this into the equation yields:

$$\Gamma = \frac{3-1}{3+1} = \frac{2}{4} = 0.5$$

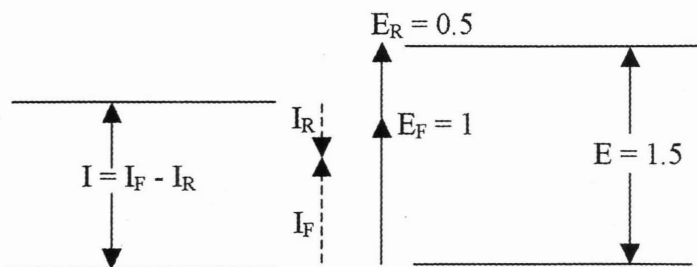


Figure 2.

Using the same forward and reflected power we can calculate the SWR from the formula below, where S is SWR and r is reflected-to-forward power ratio.

$$S = \frac{1 + \sqrt{r}}{1 - \sqrt{r}} = \frac{1 + \sqrt{0.25}}{1 - \sqrt{0.25}} = \frac{1 + 0.5}{1 - 0.5} = \frac{1.5}{0.5} = 3$$



The SWR can also be calculated directly from the reflection coefficient as shown in the following formula.

$$S = \frac{1+\Gamma}{1-\Gamma} = \frac{1+0.5}{1-0.5} = \frac{1.5}{0.5} = 3$$

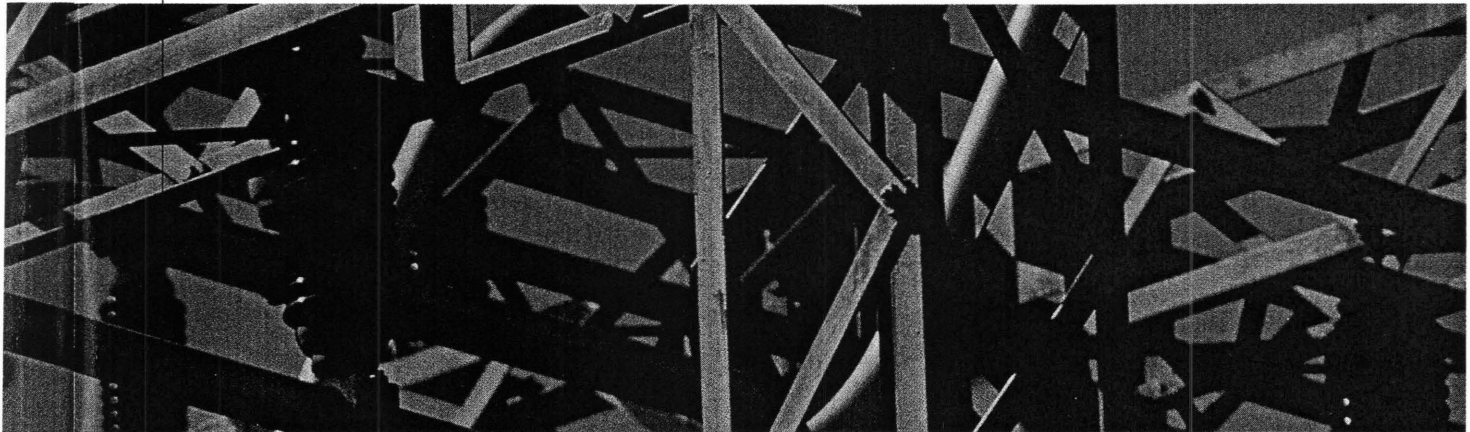
The return loss can be calculated from the reflected-to-forward power ratio from

the formula below, where LR is return loss in decibels and r is the reflected-to-forward power ratio.

$$L_R = 10 \log(r) = 10 \log(0.25) = -6dB$$

In terms of SWR, the return loss (LR) can be calculated as shown below, where S represents SWR.

$$L_R = 20 \log\left(\frac{S-1}{S+1}\right) = 20 \log\left(\frac{3-1}{3+1}\right) = 20 \log\left(\frac{2}{4}\right) = 20 \log(0.5) = -6dB$$



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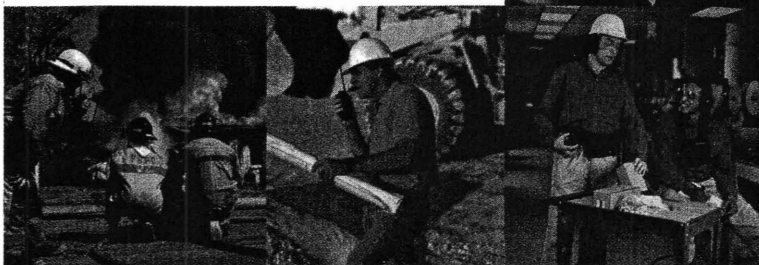
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In terms of reflection coefficient (Γ), the return loss (L_R) can be calculated as shown in the following formula.

$$L_R = 20 \log(\Gamma) = 20 \log(0.5) = -6dB$$

Figure 2 shows the current and voltage phasors for the mismatch condition discussed so far, where the forward power is 100W and the reflected power is 25W. Here, the voltage is normalized to 1 and the maximum voltage will occur when the voltage of the reflected wave is in phase with the voltage of the forward wave. The phasors representing the forward and reflected voltages rotate in opposite directions. At one point, the phasors are in phase. This produces a point of maximum voltage on the line. At another point, the phasors are 180° out of phase. This produces a point of minimum voltage on the line.

Notice, too, that when the voltage phasors combine in phase to produce a maximum voltage the current phasors are 180° out of phase, producing a point of minimum current. Therefore, this is a high impedance point on the transmission line. The normalized voltage ratio at the point of maximum voltage is equal to the reflection coefficient plus one ($1 + \Gamma$).

The maximum voltage on the line can be found by multiplying the voltage in the forward wave by the sum of $1 + \Gamma$ or 1.5. The voltage in the forward wave can be calculated from the formula below, where E is the rms voltage, Z is the system impedance and P is the power in the forward wave.

$$L_R = 20 \log(\Gamma) = 20 \log(0.5) = -6dB$$

With a forward power of 100W the voltage in the forward wave is 70.7V. With the reflection coefficient of 0.5, the maximum rms voltage on the line will be 1.5×70.7 or 106 volts rms. The peak voltage is 1.414×106 or 150 volts.

Also, the maximum rms voltage on the line can be calculated from the following formula, where S is SWR, P is net power and Z is system impedance.

$$E_{\max} = \sqrt{SPZ}$$

For example, the net power is equal to forward power (100W) minus the reflect-

ed power (25W) or 75W. Substituting this figure and an SWR of 3 into the formula yields

$$E_{\max} = \sqrt{3 \times 75 \times 50} = \sqrt{11250} = 106V$$

Again, the peak voltage can be found by multiplying the maximum rms voltage by 1.414.

Let's look at a couple of examples of coaxial cable with high VSWR. Suppose a piece of RG58 with foam dielectric is connected to a transmitter that has a power output of 500W. This cable can handle a maximum rms voltage of 600 volts. If the cable is connected to a perfectly matched load impedance, the normal rms RF voltage on the cable is found as shown below.

$$E = \sqrt{500 \times 50} = \sqrt{25000} = 158V$$

This is well within the capability of the cable. Now, suppose the load is mismatched resulting in a VSWR of 3:1 at the transmitter. Remember, if we use the VSWR to determine the maximum voltage we must calculate the net power (forward minus reflected). For a VSWR of 3:1 the net power is 375W. The rms voltage on the line is now:

$$E_{\max} = \sqrt{3 \times 375 \times 50} = \sqrt{56250} = 237V$$

According to the specification, this voltage is still well within the capability of the cable.

Now, suppose that the VSWR is 10:1. This time we will use the reflection coefficient (Γ) to determine the maximum rms voltage on the line so that we don't have to calculate the net power. In the case of a VSWR of 10:1, the reflection coefficient is 0.82. The maximum rms voltage will be:

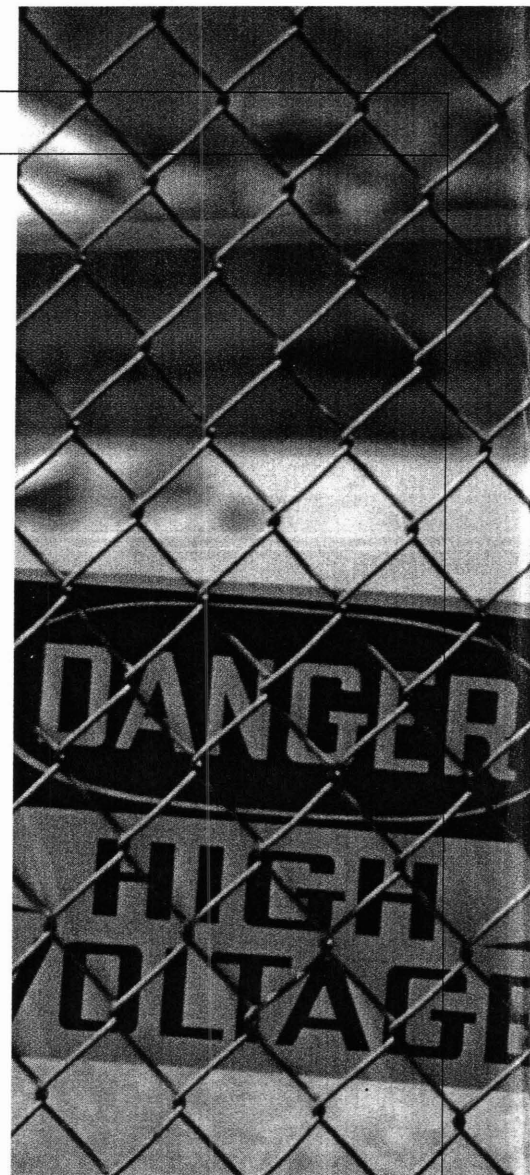
$$E_{\max} = 1.82 \sqrt{500 \times 50} = 288V$$

You can see that as the VSWR edges upward the voltage on the transmission line increases significantly. The transmitter output stage will probably sustain damage before the cable punches through. Most transmitters have fold-back circuits that serve to protect the output stage(s) from high VSWR. The power is simply reduced to a safe level until the transmitter sees the

proper load impedance. However, punch-through can and does occur.

Technically speaking, the exact point on the line where the maximum voltage occurs can be calculated if the complex impedance ($R \pm jX$) is known. However, this is usually not known to the land mobile radio technician, since measurements are usually taken with a directional wattmeter inserted into the transmission line. Practically speaking, the point of maximum voltage will always be near the transmitter end of the line. This is because of line loss and the fact that the voltage in the forward wave is always greater than the voltage in the reflected wave.

As the transmitter output power increases, more attention must be paid to a proper match at the transmitter. Losses can become quite significant and mismatches quite damaging. In the next column, we will discuss transmission line loss and its effects. Until next time-stay tuned! ■



Paging technology: POCSAG, Golay paging codes

Part 2—Although POCSAG paging code remains widely used, Golay sequential code is no longer used in most new system designs. Here are some of the advantages and disadvantages of these two early paging protocols.

Table 1—Comparison of digital paging formats.

	POCSAG	Golay	Flex	ERMES
Number of users per frequency	2 million	1 million	2 million local 1 billion national	4 million local
Frequency channel scanning	no	no	yes	yes
Data rates (bps)	512; 1,200; 2,400	300	1,600; 3,200; 6,400	6,250
Modulation type	2-level FSK*	2-level FSK	2-level FSK (at 1,600 or 3,200); 4-level FSK (at 3,200 or 6,400)	4-level FSK
Timing	synchronous	asynchronous	synchronous	synchronous
Battery-saving option	yes	yes	yes	yes
Roaming	no	no	yes, national	yes, int'l

*Frequency-shift keying

By Lynne Stewart

Table 1 at the left compares some main characteristics of four digital paging formats: POCSAG, Golay sequential code, Flex protocol, and ERMES.* Probably the most important reasons for paging system operators to use digital instead of analog formats and for the creation of the newer digital formats, Flex protocol and ERMES, is the ability to send more information, faster, to more people.

The POCSAG digital paging scheme was developed by British Telecom as a standard signaling format in the United Kingdom. It is now a widely accepted format worldwide. Official international recognition of this code was given in February 1981 when it was accepted by the International Radio Consultative Committee (CCIR, according to the French abbreviation) as the recommended Radiopaging Code No. 1 (RPC1). The CCIR is a body appointed by the United Nations to recommend worldwide standards in radio matters.

POCSAG is a synchronous paging scheme. Messages are always sent in batches, and each pager address is located in a specific frame within a batch. Figure 1 at the left shows the POCSAG paging coding format.

*POCSAG stands for Post Office Code Standardisation Advisory Group.

Golay sequential code is named after Marcel Jules Edouard Golay (b. 1902), a physicist who constructed the code about 1955.

Flex is short for flexible wide-area synchronous protocol.

ERMES stands for European radio message system.

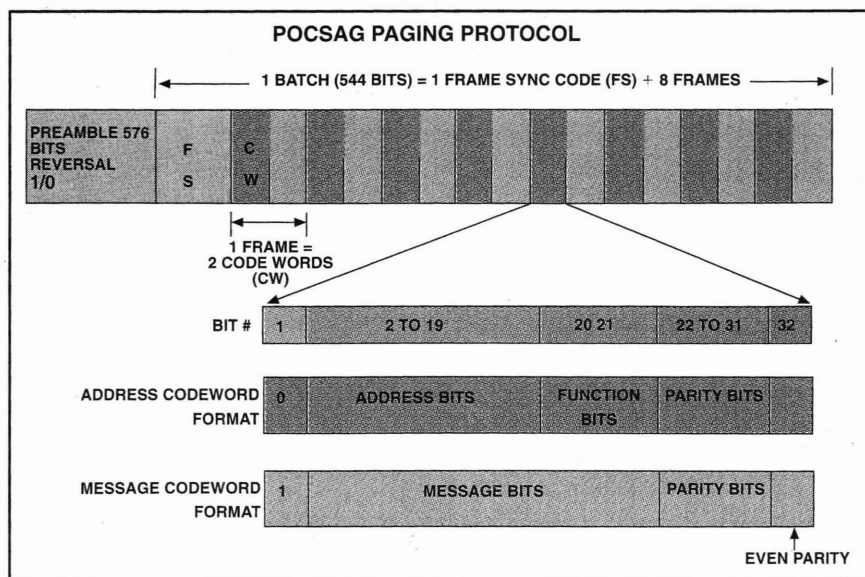


Figure 1. POCSAG is a synchronous paging scheme. Messages are always sent in batches, and each pager address is located in a specific frame within a batch.

Stewart is a market development manager with Hewlett-Packard's Microwave Instruments Division, Santa Rosa, CA.

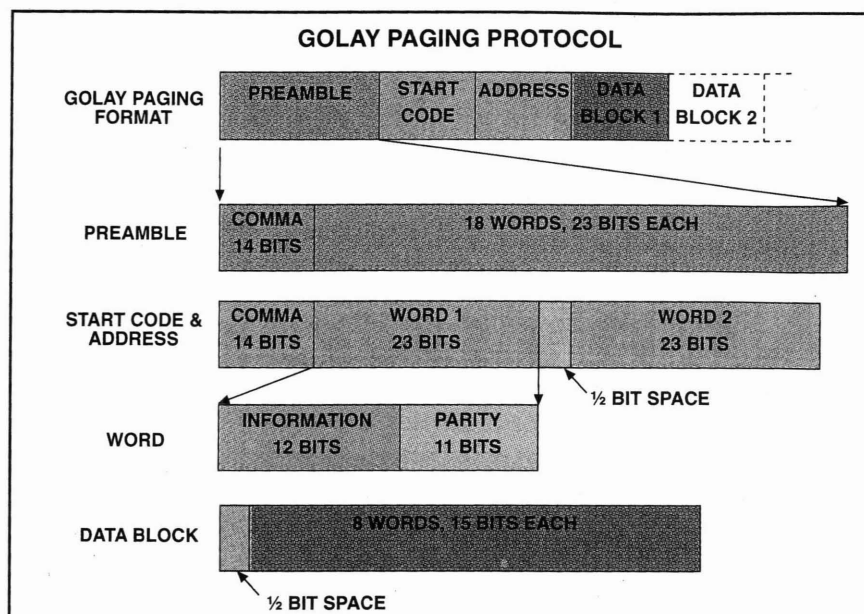


Figure 2. Golay sequential coding (GSC) is an asynchronous paging scheme. Messages are sent individually or in batches. A pager can receive its address individually or anywhere in a batch.

Battery energy consumption is reduced ("battery-saving") because each pager, after detecting the preamble and synchronizing to the frame sync word, turns off its receiver circuits until the proper frame appears. When it does look for its address in the appropriate frame, it will find it unless:

- there is no page, so an idle codeword is found.
- a message codeword from another

subscriber's page is using the frame.

In the latter case, if a message is intended for your pager, its page will be delayed until a future batch.

The number of characters that can be sent in a message codeword are 5.0 numeric or 2.8 alpha. That limit means that for a numeric pager to receive a caller's 10-digit phone number, two message codewords must be sent following the pager's address codeword.

POCSAG pagers support tone, numeric and alphanumeric messages, but not voice.

Golay sequential coding (GSC) is an asynchronous paging scheme. Messages are sent individually or in batches. A pager can receive its address individually or anywhere in a batch. Figure 2 at the left shows the Golay paging coding format.

For Golay, battery-saving is achieved through use of the preamble. There are 10 preambles, and the pagers are divided into groups based on the preamble. The pager samples every 1.38 seconds for 250ms looking for the preamble, which lasts 1.42 seconds. If its assigned preamble is found, its receiver circuitry stays on; otherwise, it returns to battery-saving mode. The polarity of the preamble signifies whether an individual or a batch call is due. Following the preamble is the start code, which marks the end of the preamble and supplies timing information to the pager.

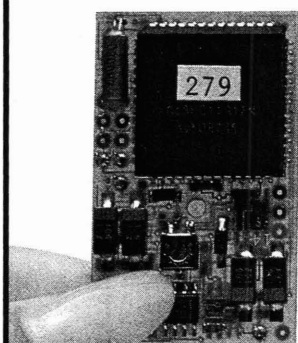
Within the address codeword, word 1 has 50 possible values, and word 2 has about 2,000 possible values, so the number of possible address codes is 100,000. Because the 10 preambles are also used to divide the pagers into groups, the maximum code capacity of 1 million is reached.

Golay pagers support tone, voice, numeric, and alphanumeric messages.

In some areas, there is a growing demand for increased capacity because existing systems are reaching their maximum capacity. Also there is a demand for more sophisticated messaging capability.

Next month: How Motorola's flexible wide-area synchronous protocol (Flex) paging format and the European Radio Message System (ERMES) paging format respond to capacity and advanced messaging capability demands.

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Circle (24) on Fast Fact Card

SWR by any other name...

By Harold Kinley, C.E.T.

No matter what you call it, SWR, or *standing wave ratio*, is an indication of the degree of match, or mismatch, of load impedance to source impedance. Generally, we use this term to indicate how well the antenna and transmission line are matched to the transmitter output. Other terms, such as forward and reflected power, reflection coefficient and return loss, are used to indicate the degree of match or mismatch as well. How are these terms interrelated? This column will explain that. Because most land mobile radio technicians are more familiar with forward and reflected power measurements, the definitions of the other terms will be based on their relationship to forward and reflected power.

Forward and reflected power

Those who work in the land mobile radio industry typically use a directional, in-line wattmeter to measure the forward power and the reflected power at the output of a transmitter. The comparison of the forward and reflected power indicates how well the load (transmission line and antenna) is accepting the power being delivered to it by the transmitter. The higher the reflected power, the higher the degree of mismatch and the lower the efficiency of the system. Not only is the mismatch reducing efficiency, but the reflected power also triggers a *foldback protection* circuit that reduces the transmitter output power. Thus, the reflected power has a two-fold effect in reducing system efficiency.

Generally, the experienced technician uses the forward and reflected power measurements to place a *qualitative* value on the antenna system. One doesn't normally think in terms of standing wave ratio, return loss or reflection coefficient. For example, a technician may measure the forward and reflected power at the transmitter output and find that the forward power is 100W and that the reflected power is 2W. This is a high ratio of forward-to-reflected power, and so all seems to be well. The qualitative analysis has been done simply by reading the forward and reflected power and mentally noting that the forward-to-reflected power ratio is high. Forget the *quantitative* analysis, right? Yes, *maybe!*

Effects of line loss

In the above example, the forward-to-reflected power ratio was 100/2, or 50:1.

Assuming a negligible line loss, this represents an excellent antenna match. However, suppose the line loss is 3dB. The line loss affects the forward power and the reflected power because both must travel the length of the line. Now, again suppose the forward power at the transmitter is 100W. With a 3dB line loss, the forward power at the antenna connection is 50W. Moreover, with a reflected power reading of 2W at the transmitter, the reflected power at the antenna must be 4W. Thus, the forward-to-reflected power at

Transmission line loss can mask the true SWR when it is measured at the transmitter vs. measurements at the antenna..

the antenna connection is actually 50:4 or 12.5:1—much worse than the 50:1 ratio found at the transmitter output. This illustrates just how transmission line loss can mask the true SWR when it is measured at the transmitter vs. measurements at the antenna. As long as you know the line loss, you can determine the true SWR at the antenna from measurements made at the transmitter. References to the previous example will be made throughout the remainder of this column.

Return loss

If the forward and reflected power are known, the return loss can be easily determined from the formula:

$$R_l = 10 \log \frac{f}{r}$$

where R_l is return loss, f is forward power, and r is reflected power.

In the example above, the forward power measured at the transmitter was 100W, and the reflected power was 2W for an f/r ratio of 50:1. The log of 50 is about 1.7. Thus, the return loss is 10×1.7 , or 17dB. Remember too, that the forward-to-reflected power ratio at the antenna was 12.5. Because the log of 12.5 is about 1.1, the return loss at the antenna is 10×1.1 , or 11dB. Notice that the difference between the return loss at the transmitter and antenna is 6dB—exactly *twice*

the line loss. The general rule can be stated as: *the return loss at the antenna will be reduced from the return loss at the transmitter by a factor of twice the line loss*. It is important to remember that *the higher the return loss, the better the impedance match*.

SWR

To convert forward and reflected power measurements into an equivalent SWR figure, the following formula is used:

$$S = \frac{1 + \sqrt{\frac{r}{f}}}{1 - \sqrt{\frac{r}{f}}}$$

where S is SWR, r is reflected power and f is forward power.

In the above example, the reflected power at the transmitter is 2W, and the forward power at the transmitter is 100W, so the ratio of r/f is 2/100 or 0.02. The square root of 0.02 is about 0.14. Thus, the SWR at the transmitter is:

$$S = \frac{1 + 0.14}{1 - 0.14} = \frac{1.14}{0.86} = 1.33$$

The reflected power at the antenna is 4W, and the forward power at the antenna is 50W. The SWR at the antenna is found using the same formula and is approximately equal to 1.79. Again, the line loss masks the true SWR at the antenna as seen from the transmitter end.

Reflection coefficient

The reflection coefficient can be derived from forward and reflected power measurements by using the following formula:

$$\Gamma = \sqrt{\frac{r}{f}}$$

where Γ represents reflection coefficient, r is reflected power and f is forward power.

Using the example at the transmitter, the ratio of r/f would be 2/100, or 0.02.

(continued on page 50)

Kinley, a certified electronics technician, is regional communications manager, South Carolina Forestry Commission, Spartanburg, SC. He is a member of the Radio Club of America. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, which is available for direct purchase. Write to 204 Tanglewylde Drive, Spartanburg, SC 29301. Kinley's email address is hkinley@aol.com.

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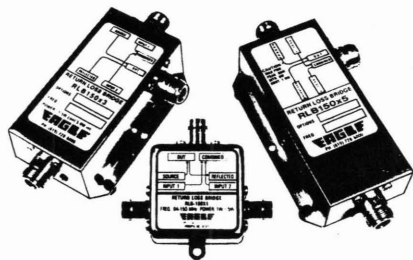
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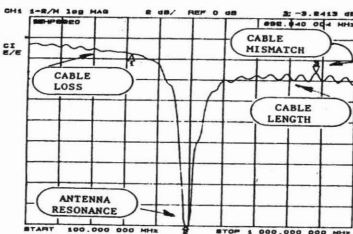
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Technically speaking

(continued from page 8)

The square root of 0.02 is approximately equal to 0.14. Thus, the reflection coefficient at the transmitter is 0.14. At the antenna, the ratio of r/f is 4/50, or 0.08. The square root of 0.08 is approximately equal to 0.28. It is interesting to note that for a line loss of 3dB, the reflection coefficient at the antenna is exactly *twice* the reflection coefficient at the transmitter.

Transmission loss

Transmission loss is a term used to describe just how much the signal power is reduced by a given mismatch, zero being the perfect figure for a *return loss* of ∞ dB, or a *standing wave ratio* of 1:1 or a *reflection coefficient* of 0.0. The transmission loss can be determined from forward and reflected power by using the following formula:

$$T_L = 10 \log \left(1 - \frac{r}{f} \right)$$

where T_L is transmission loss, r is reflected power and f is forward power.

Using our example, the forward power at the antenna was 50W, and the reflected power was 4W. Thus, the ratio r/f is 4/50, or 0.08, and $1 - 0.08 = 0.92$. The log of 0.92 is approximately equal to 0.036, and 10×0.036 is 0.36dB. Therefore, the transmission loss caused by the mismatch at the antenna is 0.36dB.

Additional line loss

An additional line loss is caused by the standing waves on the line. The net power absorbed or radiated by a load is equal to the forward power minus the reflected power. The loss of the line can be calculated from the net input power minus the net output power. In our example, the net input power was 98W, and the net output power was 46W. Calculating the loss from input to output results in a figure of about 3.28dB. This is 0.28dB more than the loss of the line under normal matched conditions. The additional 0.28dB of loss is caused by the standing waves.

Summary

No matter what you might call it, the results are the same. A mismatched load will cause degradation in transmission efficiency. However, the actual transmission loss is not as severe as one might expect. Transmission line loss will mask the true SWR that exists at the antenna when measurements are made at the transmitter. By using the formulas presented here you can convert forward and reflected power into any of the other units of measure.

Until next time—*stay tuned!*



Field-tuning isolators

By Harold Kinley, CET

The June and July columns covered ferrite isolators in some depth.

It is time to put the subject aside for a while—right after these few tips about field-tuning or retuning isolators.

The first and probably the best advice to give about the field-tuning of isolators is to follow the manufacturer's instructions to the letter. After all, the manufacturer knows the limitations and capabilities of his isolator better than anyone else.

General rules:

If possible, an isolator should be tuned in the installed position.

Generally, isolators should not be installed on a steel (or other ferrous metal) surface, but some manufacturers state that it is okay to do so as long as the tuning is "touched up" after installation.

If such isolators were tuned on a workbench and then installed on a steel

surface, it is likely that some detuning would result. The severity of the detuning would depend upon the degree of magnetic shielding the case provides.

Other manufacturers claim that their isolators are so well shielded that the mounting surface has a negligible effect on isolator tuning. Again, follow the manufacturer's instructions and use common sense.

Signal source

If you use the transmitter output as the signal source for tuning the isolator, reduce the transmitter power output to about 10% of the normal isolator power input rating (not to exceed 10W) while tuning the isolator. Once the isolator is tuned, normal rated input power can be applied to the isolator.

There are many accepted methods by which an isolator can be properly tuned. These methods range from very simple to fairly complex. The method you use will depend on the type of equipment you have available and just how much of a "purist" you are.

The methods presented here are not a complete compilation of all the possible ways to tune an isolator. There are probably at least 101 ways!

► **Single-section isolator**—The single-section isolator is the simplest to tune.

First, remove the caps to gain access to the tuning adjustments. A single-section isolator has three adjustments: *input port*, *output port* and *load port*.

The following procedure uses an in-line directional wattmeter as the measuring instrument and the transmitter as the signal source. (See Figure 1 below.)

(1) Figure 1A. Key the transmitter (low power) and tune the input port (1) and output port (2) for maximum power indication on the wattmeter.

(2) Figure 1B. Reverse the connections to ports 1 and 2. Key the transmitter and adjust the port 3 trimmer for minimum power indication.

► **Dual-section isolator**—The following system is used for numbering the various tuning adjustments for the dual isolator discussed in these procedures.

- 1 = input port of first section.
- 2 = output port of first section.
- 3 = input port of second section.
- 4 = output port of second section.
- 5 = load port of first section.
- 6 = load port of second section.

Procedure 1: This procedure uses an in-line directional wattmeter as the measuring instrument and the transmitter as the signal source. Refer to Figure 2 on page 68 for this procedure.

(1) With the setup shown at Figure 2A, key the transmitter and adjust trimmers 1, 2, 3 and 4 for maximum power indication. Repeat this adjustment sequence until no further improvement can be made.

(2) Using the setup at Figure 2B, key the transmitter and adjust trimmer 5 (load port of first section) for minimum power indication.

(3) Using the setup at Figure 2C, key the transmitter and adjust trimmer 6 for minimum power indication.

(4) With the test setup at 2D and with the transmitter keyed, there should be no discernible indication on the wattmeter.

(5) Measure the insertion loss and the isolation. If both are not within the manufacturer's specifications, repeat
(continued on page 68)

Kinley is a certified electronics technician with the South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985.

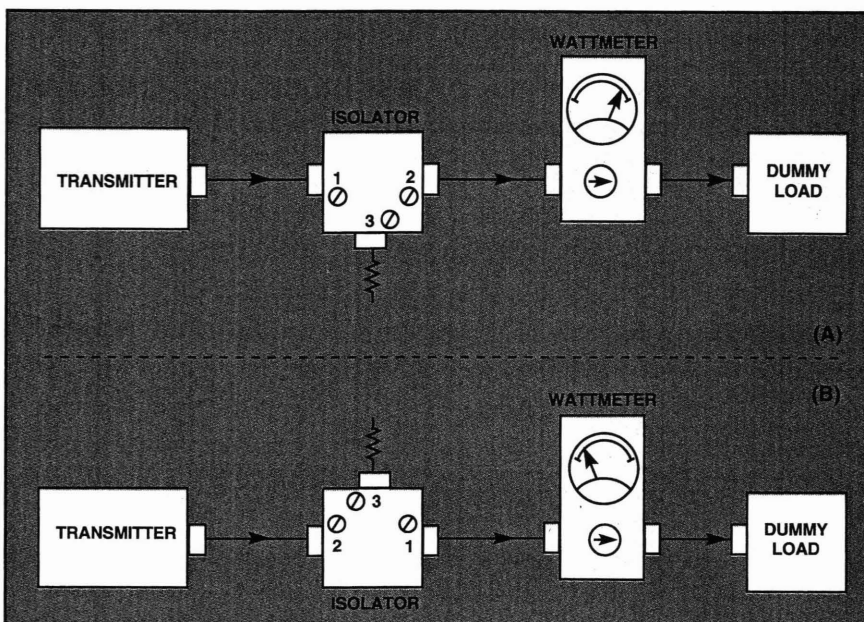


Figure 1. The single-section isolator is the simplest to tune. This diagram shows the equipment configurations to use with the two-step tuning procedure described in the text.

Technically speaking

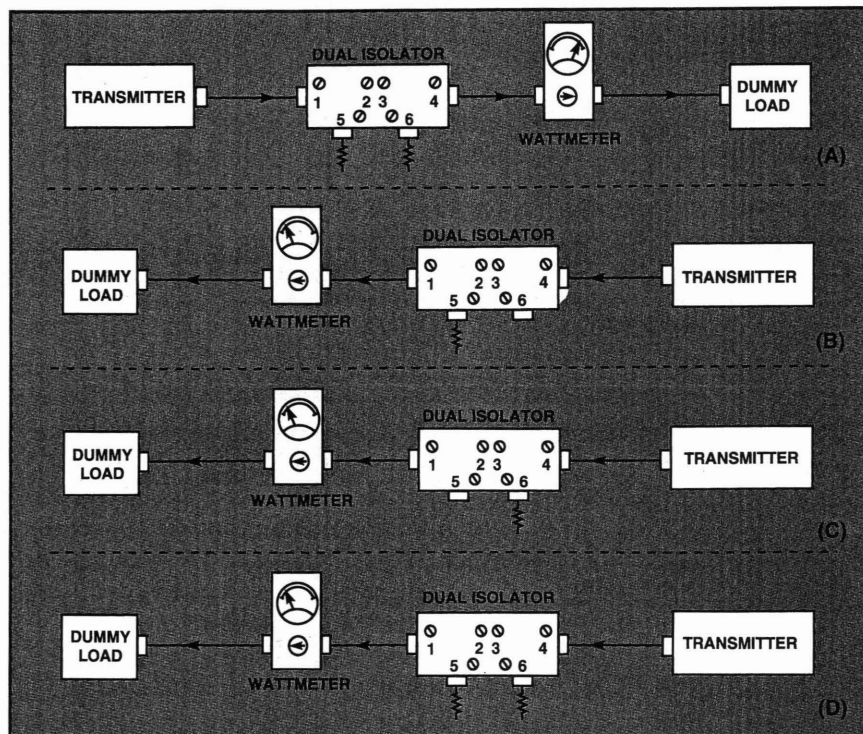


Figure 2. The first of three procedures for tuning dual-section isolators described in the text uses equipment configurations shown above.

(continued from page 8)

the tuning procedure.

Procedure 2: In this setup, a signal generator (highly stable) and a spectrum analyzer are used in place of the transmitter and wattmeter as the signal source and measurement instruments.

The signal generator frequency is set to the frequency to which the isolator is to be tuned. The spectrum analyzer is tuned to the same frequency.

If the spectrum analyzer is part of a service monitor, the frequency can be monitored on the frequency readout to keep the signal generator on frequency. If the signal generator is synthesized, you will not have to worry about frequency drift. Refer to Figure 3 on page 70 for this procedure.

(1) At Figure 3A, trimmers 1, 2, 3 & 4 are tuned for maximum indication on the spectrum analyzer display.

(2) At Figure 3B, trimmer 6 is adjusted for minimum indication on the spectrum analyzer display.

(3) At Figure 3C, trimmer 5 is ad-

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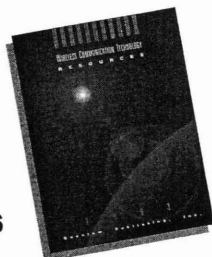


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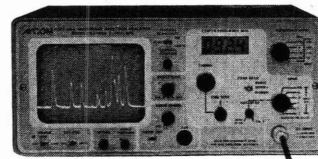
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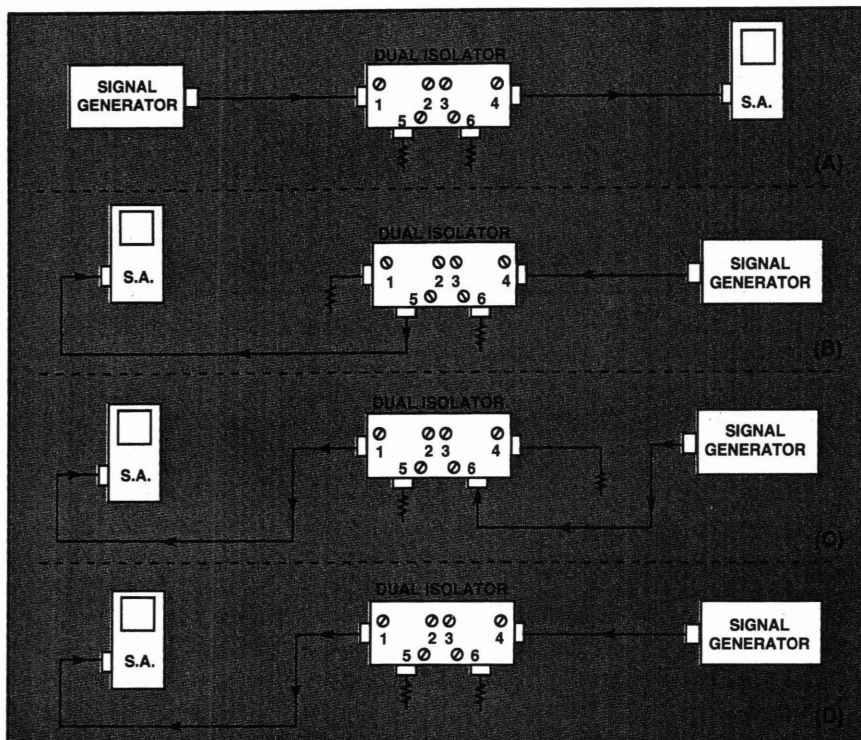


Figure 3. The second of three procedures for tuning dual-section isolators described in the text uses equipment configurations shown above.

justed for minimum indication on the spectrum analyzer display.

At Figure 3D, trimmers 5 and 6 are adjusted for minimum signal level on the display.

Procedure 3: The equipment required are a return loss bridge and spectrum analyzer-tracking generator combination. Refer to Figure 4 on page 72 for the following steps.

(1) At Figure 4A, trimmer 1 is adjusted for maximum return loss on the spectrum analyzer display.

(2) At Figure 4B, trimmer 4 is adjusted for maximum return loss on the spectrum analyzer display.

(3) At Figure 4C, trimmers 2 and 3 are adjusted for the best symmetry of the passband response of the isolator with the peak of the response curve at the desired center frequency.

(4) At Figure 4D, trimmer 5 is adjusted for minimum signal level on the display.

(5) At Figure 4E, trimmer 6 is adjusted for minimum signal level on the display.

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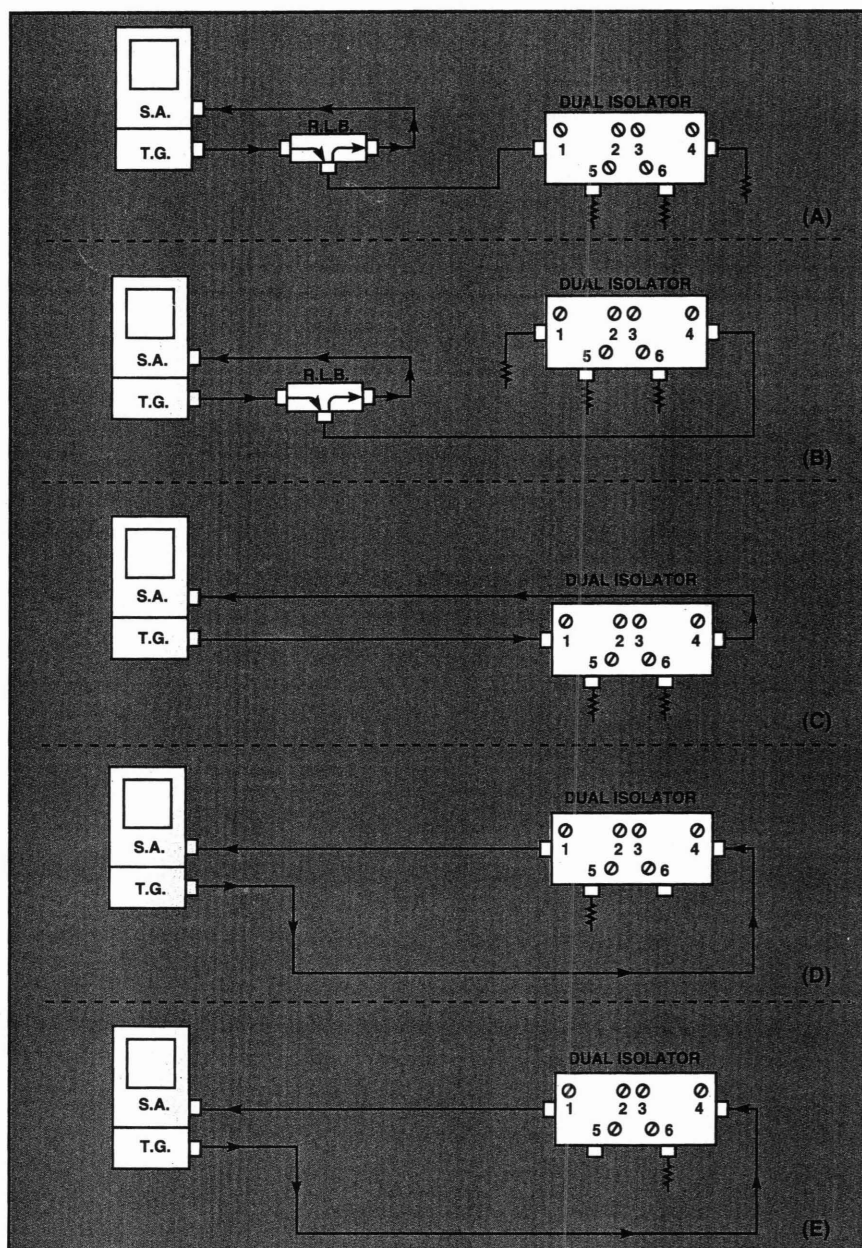


Figure 4. The third of three procedures for tuning dual-section isolators described in the text uses equipment configurations shown above.

Reconnect a load to port 5, and adjust trimmers 5 and 6 for minimum signal level on the display.

When tuning a dual isolator, never tune the load port trimmer with the associated load resistor removed. Remove one load at a time and adjust the opposite load port trimmer for maximum isolation (minimum signal level in the reverse direction).

After retuning an isolator, always check the insertion loss (forward direction) and isolation (reverse direction) and compare the results with the

manufacturer's specifications. If the measurements do not meet the specification, retune the isolator. If the isolator still does not come up to specifications, consult the manufacturer.

As stated, these procedures are not all-inclusive. There are many other variations that are correct and proper that will get the job done.

The procedure that you use will depend on you and your particular arsenal of test equipment. Again, use common sense, and consult the manufacturer when you are in doubt.

How to use duplexers: Isolation requirements

Part 1—Duplexers in various forms isolate simultaneously operated transmitters and receivers. Here are some steps to follow to calculate the required isolation. Impedance differences make duplexers work.

By Brian J. Henderson, P.E.

You have just installed a base station and mobile radio system, perhaps for a customer.

Coverage is satisfactory from the base station to the mobile radios. Mobile-to-mobile range is limited. The customer wants to consider a repeater system to improve mobile range.

You want to serve the customer the best way you can. You buy a repeater. You install an antenna. You know you must buy a duplexer. Why? What does a duplexer do?

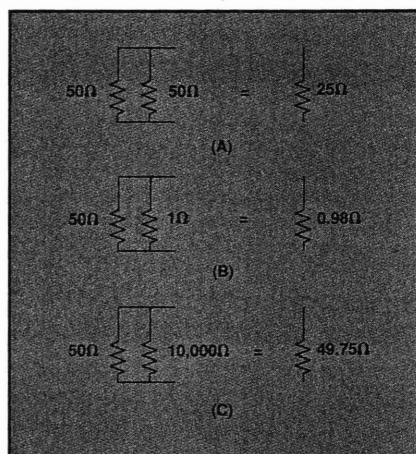


Figure 1. A duplexer uses impedance differences to isolate the transmitter and receiver. At (A), two 50Ω resistors or 50Ω coaxial cables wired in parallel have a resulting impedance of 25Ω . When connected to another 50Ω device, the connection represents an impedance mismatch and results in a power loss. At (B), when a 50Ω impedance circuit is connected in parallel with a 1Ω impedance circuit, the combined impedance is extremely close to 1Ω , almost a short circuit. At (C), when a 50Ω impedance circuit is connected in parallel with a $10,000\Omega$ impedance circuit, the combined impedance is extremely close to 50Ω , meaning that there is little effect upon the 50Ω circuit's impedance.

A duplexer allows the use of a single antenna for transmitting and receiving simultaneously, as a repeater usually does.

Duplexers come in many shapes, sizes and types from many manufacturers. They often are described in antenna catalogs. Duplexers also may be called filters, combiners and cavities. Many people consider them to be mystical things that only engineers know about.

Duplexers, including filters, reject, pass and other "cavity devices," actually are quite straightforward once you understand the basics. The following describes the basics of filters, reject, pass and other "cavity filter devices."

Why is a duplexer required?

There are three reasons why a duplexer is used.

The first is to provide isolation between the transmitter and receiver.

The second is to preserve impedance matches among the antenna, transmitter and receiver. Impedance must be maintained at 50Ω to keep everything operating at maximum efficiency.

The third is to prevent several transmitters and receivers from interfering with each other and causing or receiving intermodulation products and other interference.

Isolation

In a mobile or base station radio, the transmitter and receiver do not operate at the same time.

When the receiver is on, and the channel is being monitored, the transmitter is turned off. When the operator presses the push-to-talk (PTT) button and transmits, the receiver is turned off.

A repeater receives a signal in the receiver and rebroadcasts the audio through the transmitter at the same time. This type of operation is called full-duplex because

the transmitter and receiver are both on and operating at the same time. Full-duplex operation gives its name to the "duplexer."

Note that the receiver is receiving a signal of only a few microvolts (extremely small). The transmitter is transmitting a signal of several volts (extremely large). What is required is a device to separate or to isolate the receiver from the transmitter so that the two can operate at the same time without interfering with each other.

The receiver must continue to receive extremely small signals while the transmitter transmits extremely large signals. That is where the duplexer comes in. It will provide the required isolation between transmit and receive frequencies.

As an example of isolation, consider a 25W repeater package. The transmitter operates at a power level of 25W or +44.0dBm. The receiver is rated at $0.5\mu V$ or -113.0dBm. The difference between these two signal levels is 157dB.

The repeater must operate at both of these power levels. Using two different frequencies provides a certain amount of isolation. A duplexer provides additional isolation so that the repeater can operate in full-duplex mode.

A duplexer prevents receiver desensitization (desense) from occurring. That means that the receiver is as sensitive when the transmitter is on as when the transmitter is off.

How much isolation?

How much isolation is necessary depends on two factors: the amount of frequency separation between transmitter and receiver, and the transmitter output power.

Frequency separation is the main

Henderson is senior engineer, communications, with Canadian Western Natural Gas, Calgary, Alberta.

governing factor. Manufacturers graph *duplex curves* for each radio they produce. These curves tell you how much transmitter noise is produced by a transmitter at a given receiver separation from the transmit frequency. Sometimes these curves are available from the manufacturer; sometimes they are not.

The manufacturer always specifies *adjacent channel selectivity*. This specification can be used as a guideline. Adjacent channel selectivity is usually specified for either 25kHz or 30kHz channel spacing, and that is the spec that can be used.

Transmit power is the other factor. The more transmit power, the more isolation required. All calculations should be based on maximum transmit output power. Otherwise, turning the power up on the transmitter causes receiver desense, and the repeater will not operate properly.

As an example, take the same 25W transmitter as before.

25W = +44.0dBm
 $0.5\mu V = -(-113.0\text{dBm})$
 Receiver selectivity = -85dB

Minimum required isolation = 72.0dB
 3dB safety factor = 3.0dB

Total isolation required = 75.0dB

As a rule of thumb, when using this method, make sure that you have at least 3dB additional isolation to allow for any increase in transmitter power or decrease of receiver rejection in the future. In this case, isolation should be $72.0\text{dB} + 3.0\text{dB} = 75\text{dB}$ minimum.

If the transmitter power is increased to 50W, (+3dB) then isolation must be increased the same amount, by +3dB, to 78dB.

In this example, look for a duplexer specified as having a minimum of 75dB isolation. Note that, depending on frequency spacing, the amount of isolation required may or may not be available.

As a rule, at VHF, 300kHz is the minimum channel spacing that can be used in any repeater system. At UHF, a minimum frequency spacing of 3MHz is required.

Impedance

The duplexer uses impedance differences to isolate the transmitter and receiver.

For example, two 50Ω resistors or 50Ω coaxial cables wired in parallel have a resulting impedance of 25Ω. (See Figure 1A on page 10.) When connected to another

50Ω device, the connection represents an impedance mismatch and results in a power loss.

When a 50Ω impedance circuit is connected in parallel with a 1Ω impedance circuit, the combined impedance is extremely close to 1Ω, almost a short circuit. (See Figure 1B).

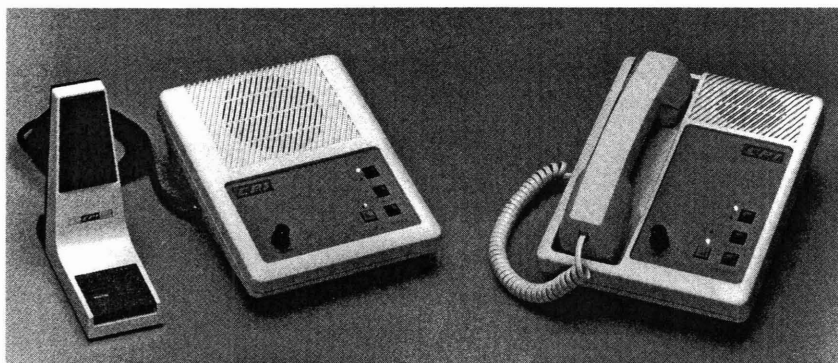
When a 50Ω impedance circuit is connected in parallel with a 10,000Ω impedance circuit, the combined impedance is extremely close to 50Ω, meaning that there is little effect upon the 50Ω circuit's impedance. (See Figure 1C.)

Duplexers are tuned devices.

They can present a 50Ω impedance at one frequency and either a lower or a higher impedance at another frequency. To provide isolation between two frequencies, a duplexer presents a short or an open circuit to a transmission line at a particular frequency. Energy at that frequency then either is removed from a transmission line or is prevented from traveling through the transmission line.

Next: Applications for types of duplexers are described to help you choose and configure cavities for specific purposes. Frequency bands and frequency separation play a part in selecting the cavities.

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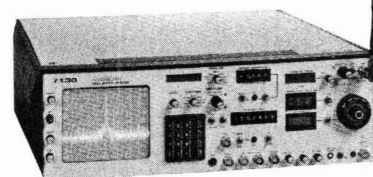
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How to use duplexers: The various types

Part 2—Applications for types of duplexers are described to help you choose and configure cavities for specific purposes. Frequency bands and frequency separation play a part in selecting the cavities.

By Brian J. Henderson, P. Eng.

Duplexers allow transmitters and receivers to operate simultaneously and to use the same antenna, as they do in typical repeater installations.

Various configurations can be used to filter interference, too.

The bandpass cavity

The bandpass cavity is the easiest type of duplexer to explain and to understand.

First, imagine a transmitter and receiver antenna separated by some distance. (See Figure 1A to the right.) The transmitter transmits energy that is radiated by the antenna and received by the receiver antenna and receiver. There is a loss between the two antennas.

If these two antennas were contained inside a cylinder, the losses between them would be reduced. Instead of the transmitting antenna radiating some energy in all directions, most of the energy would be contained inside the cylinder. Most of the energy would be picked up by the receiving antenna. Losses are reduced. (See Figure 1B).

The cylinder's length and volume affect the transmission path between the two antennas. This path between the two antennas is quite sharp and pronounced.

If the antennas are simple "loops" or short dipoles on opposite sides of a cylinder and the volume of the cylinder is adjusted by a conductive rod in the middle, the previously imaginary configuration has become a bandpass cavity filter. (See Figure 1C).

The bandpass cavity filter's frequency response is shown in Figure 2 top right.

Henderson is senior engineer, communications, with Canadian Western Natural Gas, Calgary, Alberta.

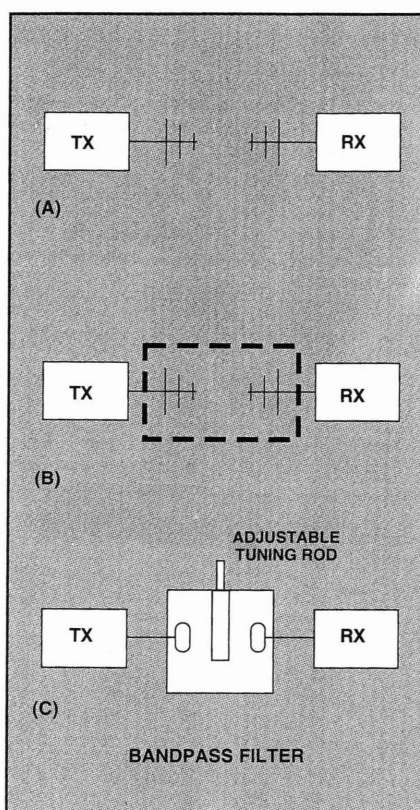


Figure 1. At (A), radio energy transferred between a transmitter antenna and receiver antenna is subject to a loss. If the antennas were within a cylinder, as in (B), the loss would be reduced. The cylinder's length and volume affect the transmission path between the two antennas. If the antennas are simple 'loops' or short dipoles on opposite sides of a cylinder, and the volume of the cylinder is adjusted by a conductive rod in the middle, the previously imaginary configuration has become a bandpass cavity filter, as in 1C.

The bandpass cavity can be adjusted for frequency by adjusting a "trombone slide" inside the cavity. Loss between the transmit and receive ports is at a minimum at the bandpass frequency and increases be-

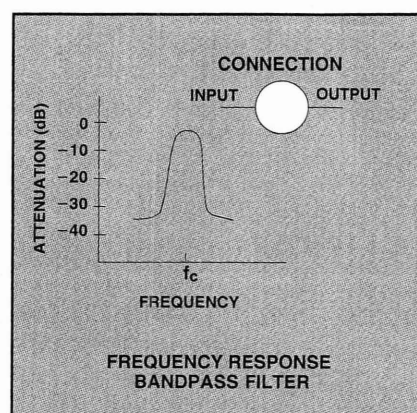


Figure 2. This is the frequency response of the bandpass cavity filter in Figure 1. The cavity can be adjusted for frequency by adjusting a "trombone slide" inside the cavity. Loss between the transmit and receive ports is at a minimum at the bandpass frequency and increases beyond the cavity's passband.

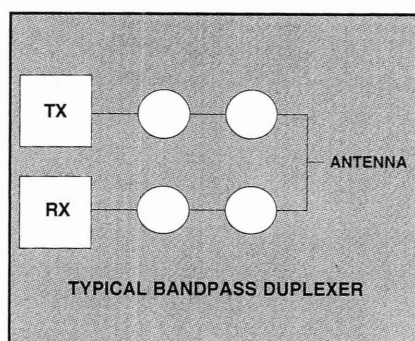


Figure 3. Several bandpass cavities can be combined to separate transmit and receive frequencies, as shown above. The idea is to provide enough isolation between a transmitter and receiver so that the same antenna can be used for both functions.

yond the cavity's passband.

In addition, the SWR is at a minimum at the tuned frequency. SWR rises outside the bandpass cavity's passband.

Note that the physical diameter of the

cavity has the effect of increasing the circuit "Q." The larger the diameter, the narrower the bandpass cavity and the higher the "Q." A 7-inch cavity is a better bandpass filter than a 5-inch cavity. A 10-inch cavity is better yet.

The bandpass duplexer

The function of a bandpass cavity has now been explained. So what does this do for the duplexer?

Several bandpass cavities can be combined to separate transmit and receive frequencies. The idea is to provide enough isolation between a transmitter and receiver so that the same antenna can be used for both functions. Such a duplexer is shown in Figure 3 on page 18.

Generally, two bandpass cavities are used in the transmit leg, and two are used in the receive leg of a repeater. Duplexers are connected in series with transmit and receive ports, as shown in Figure 3. The antenna port simply uses a "T" connector to parallel transmit and receive cables.

There is an advantage to using bandpass cavities to build a duplexer. The passbands of both the transmitter and receiver are narrowed.

In the case of the receiver, interference

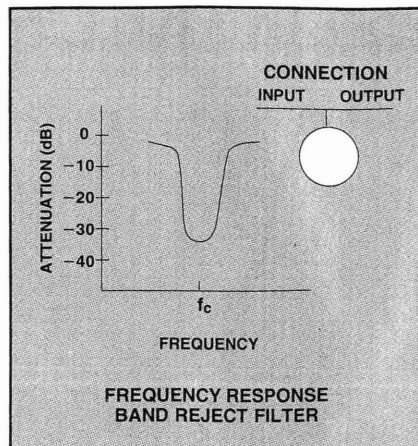


Figure 4. A reject cavity is a bandpass cavity with only one connector. The other connector either is left open or is omitted. The cavity's spectrum plot and connection diagram are shown above.

and intermodulation outside the passband of the receive cavities is reduced. Interference to other systems from your transmitter will be reduced.

There is one drawback to using bandpass cavities for a duplexer. Practical limits on construction constrain transmit-receive spacing to less than 2MHz at VHF and

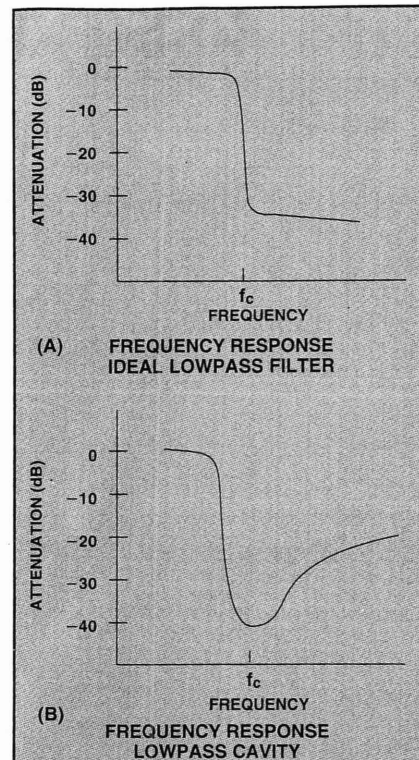


Figure 5. The graph at (A) shows the theoretical performance of a lowpass filter. It has a defined cutoff frequency. Below that frequency, signals pass with little loss. Above that frequency, signals are shorted out and do not pass through. A lowpass filter can be made by adding a capacitor in series with a bandpass cavity's feed loop. The pattern shown at (B) is that of a lowpass filter.

5MHz at UHF frequencies. Bandpass filters do not provide enough isolation for closer transmit-receive frequency spacing.

The reject cavity

Another common type of filter is called a reject cavity.

A reject cavity is simply a bandpass cavity with only one connector. The other connector either is left open or is omitted. See the spectrum plot and connection diagram shown in Figure 4 above left.

When connected into a transmission line with a "T" connector to "short out" a single frequency, the reject cavity removes the frequency from the line. This method commonly is used when another transmitter interferes with a receiver. A reject cavity is connected into the receive path to "notch out" the offending signal.

The lowpass filter

Another type of filter commonly used is the lowpass filter.

The theoretical performance of a lowpass filter is shown in Figure 5A above right. A lowpass filter has a defined cutoff frequency. Below that frequency, signals

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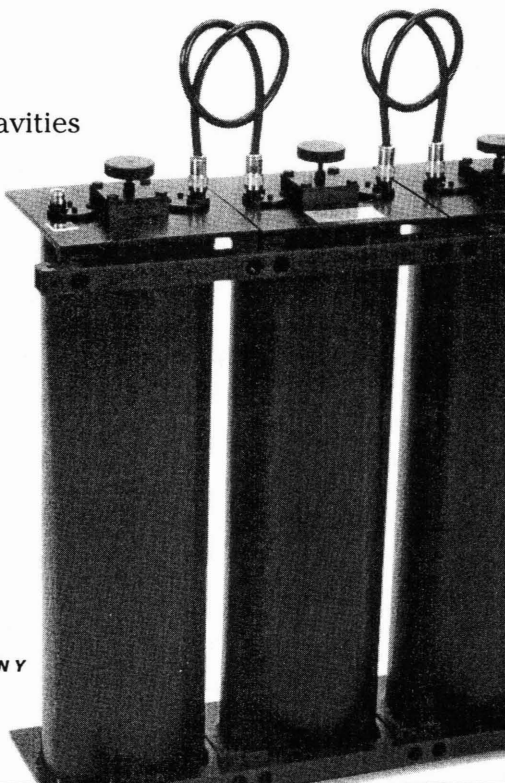
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pass with little loss. Above that frequency, signals are shorted out and do not pass through.

A lowpass filter can be made by adding a capacitor in series with a bandpass cavity's feed loop. The pattern shown in Figure 5B is that of a lowpass filter.

The highpass filter

A highpass filter is the opposite of a lowpass filter.

Signals above the cutoff frequency pass through with little loss, and signals below the cutoff frequency are shorted out and do not pass through the highpass filter. (See Figure 6A to the right.)

The highpass filter also can be made by adding a capacitor in series with a reject cavity's feed loop. The pattern shown in Figure 6B is that of a highpass filter.

Note that the highpass and lowpass filters have exactly the same construction.

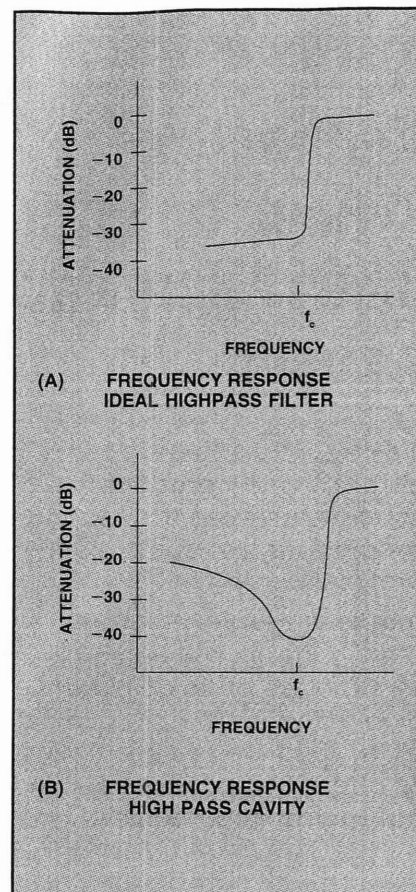


Figure 6. A highpass filter is the opposite of a lowpass filter. As shown at (A), signals above the cutoff frequency pass through with little loss, and signals below the cutoff frequency are shorted out and do not pass through the highpass filter. The highpass filter also can be made by adding a capacitor in series with a reject cavity's feed loop. The pattern shown at (B) is that of a highpass filter.

Near the feed connector on a filter you will see a small screw for adjusting the capacitor.

The capacitor setting determines whether the filter is a highpass or lowpass device, and it is not possible to tell from a visual inspection which type the cavity is. The only way to tell is by looking at the display of the cavity on a spectrum analyzer.

The bandpass/band-reject cavity

Highpass and lowpass filters could be combined together to form a duplexer.

There is a better way. A proper capacitor choice makes the filter a combination lowpass-highpass and bandpass filter, depending on the capacitor setting. (See Figures 5B and 6B.)

In this way, a single cavity connected to a receiver, for example, can pass the receive frequency with little loss and reject the transmit frequency. This forms the

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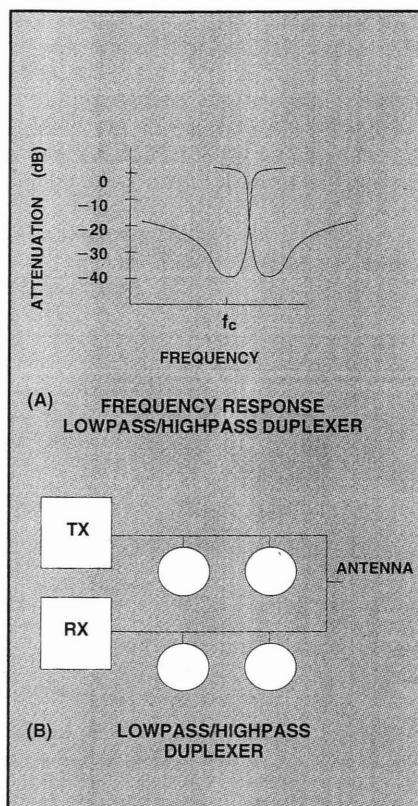
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basis of the typical highpass-lowpass duplexer.

The bandpass/band-reject duplexer

Remember the isolation calculations performed earlier?

To get the 70dB–75dB of isolation required, two bandpass/band-reject cavities can be connected into each of the repeater's transmit and receive legs. (See Figure 7 to the left.)

Two cavities in the transmit side and two cavities in the receive side provide the required 70dB–75dB of isolation. In this case, minimum frequency separation is about 500kHz.

If a frequency spacing less than 500kHz is required, three cavities can be used on each side. This configuration increases isolation to about 90dB. The minimum fre-

quency separation is about 300kHz at VHF. This separation is a practical limit of this type of duplexer. Adding more cavities increases the loss while providing little additional rejection.

Next: How duplexers are constructed and how they are installed have much to do with their tuning and frequency stability. Suggestions are included for ordering duplexers and for initial tuning or retuning.



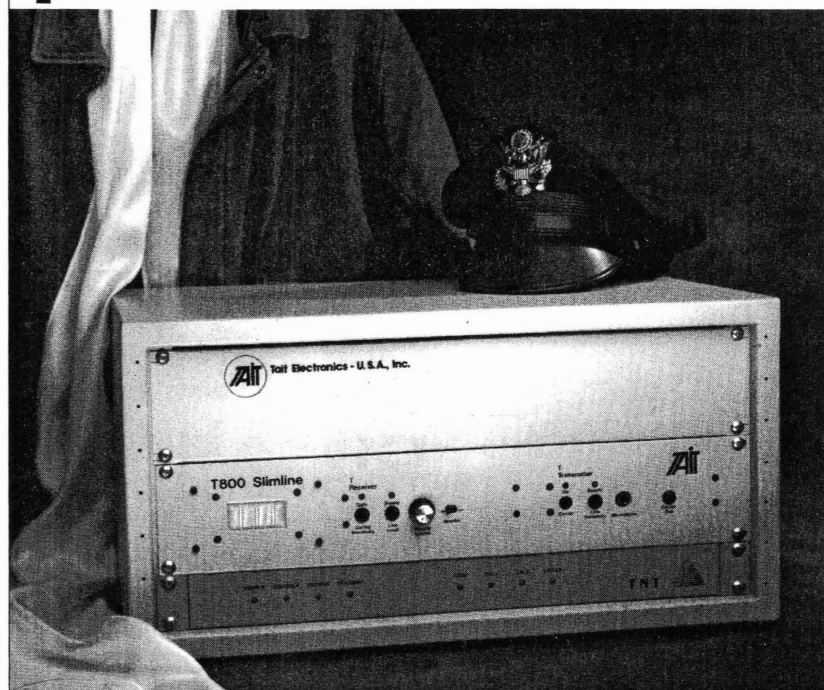
Duplexer series

The previous installment in this article series is "How to use duplexers: Isolation requirements" in the July 1994 issue.

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Figure 7. Remember the isolation calculations described in the text? To get the 70dB–75dB of isolation required, two bandpass/band-reject cavities can be connected into each of the repeater's transmit and receive legs, as shown to the left. Two cavities in the transmit side and two cavities in the receive side provide the required 70dB–75dB of isolation. In this case, minimum frequency separation is about 500kHz.

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Diversity reception controller helps radio system performance

Part 2—Using amplitude sensing, analog logic and output filtering along with low-cost integrated circuitry, an ideal selection diversity controller brings the advantages of diversity reception to commercial mobile communications.

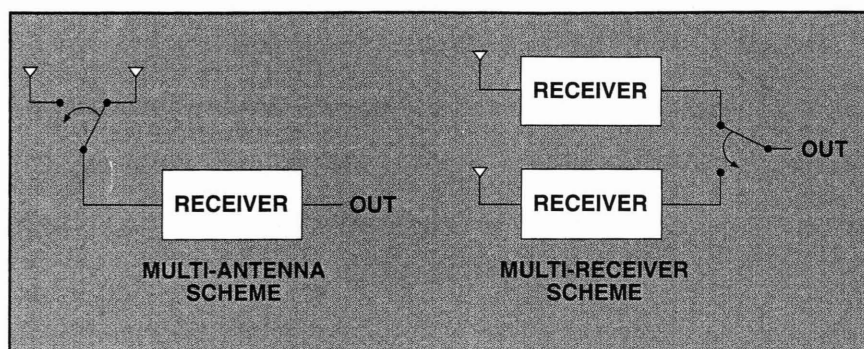


Figure 1. There are literally hundreds of variations on the concept of diversity reception. Most can be grouped into two main classifications of diversity reception systems, *multi-antenna systems* and *multi-receiver systems*, as shown above. Both of these schemes have proven largely unsuccessful, both technically and commercially, for a number of reasons.

By Shane Fitzgerald

Short-term (multipath) fading, as described in the first installment of this two-part article, is extremely destructive to high-speed data communications. What the ear would endure as one little pop or a hissing sound may destroy several hundred bits of information. An 800MHz mobile unit traveling at 15mph through a stationary environment will encounter 38 fades per second, and more if the environment includes moving objects such as trains, aircraft and automobiles. If these fades sink below the receiver's data recovery threshold, successful data reception becomes nearly impossible.

How can the integrity of data communications be protected in a fading environment? Two of the most popular and widely used methods are *forward error-correction*

Fitzgerald is RF design engineer at ElectroCom Communications Systems, Santa Fe Springs, CA. ElectroCom manufactures the ideal selection diversity controller described in this two-part article.

(FEC) and redundant transmissions.

With FEC, message reconstruction information is attached to each message prior to transmission. Should the message be received with errors, the reconstruction information is used to reconstruct the message. This method is complex. Message reconstruction taxes the receiver's processing power because it is computationally intensive, and it reduces system throughput significantly because an FEC code is added to each transmission.

With redundant transmissions, each message is transmitted multiple times and a majority voting process is used to decode the message. This simple scheme is less computationally intensive than FEC; but multiple transmissions of identical information are extremely inefficient and dramatically reduce system throughput.

FEC and redundant transmissions share two fundamental flaws in dealing with the destructive interference caused by multipath fading. The first flaw these systems exhibit is their fatalistic approach to the problem. Instead of trying to eliminate the source of the problem (short-term fading), these techniques try to recover from

the damage the fading causes. The other flaw is that if the receiver remains stationary while in a deep fade, no amount of FEC or redundant transmissions will help; communication will not be possible.

Diversity reception, on the other hand, reduces destructive effects of multipath fading directly. Two closely spaced antennas receive signals that are considered to be uncorrelated; therefore, when one antenna experiences a fade, then the probability of the other antenna simultaneously experiencing a fade is extremely unlikely. By selecting between antennas or receivers quickly enough and in response to the fades, the damaging effects of short-term fading are dramatically reduced.

There are literally hundreds of variations on the concept of diversity reception. Most can be grouped into two main classifications of diversity reception systems, *multi-antenna systems* and *multi-receiver systems*, as shown in Figure 1 at the left. Both of these schemes have proven largely unsuccessful, both technically and commercially, for a number of reasons.

Multi-antenna diversity uses a single receiver with multiple antennas that are connected to it one at a time. When the receiver picks up an adequate signal from its current antenna, it maintains the connection. If the signal falls below a predetermined threshold, then the receiver switches to another antenna that may deliver a better signal.

Unfortunately, there is no guarantee that a better signal will be found when the receiver switches antennas. The signal may, in fact, be worse. Moreover, waiting until the threshold of reliable communication is reached before switching means that a problem already has been encountered, rather than avoided.

Multi-receiver diversity uses multiple receivers to supply recovered modulation. Special circuitry determines which

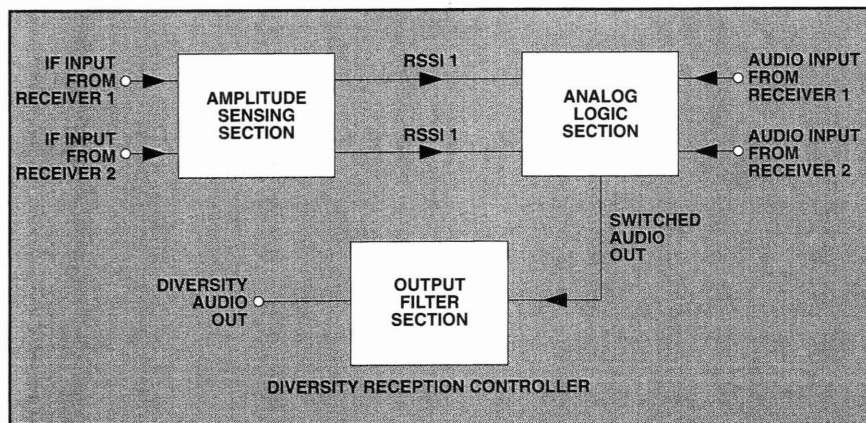


Figure 2. The diversity reception controller includes three main sections: an amplitude-sensing section, an analog logic section, and an output filter section.

receiver is receiving the better signal, and that receiver is then selected. Composite recovered modulation is assembled using a combination, or selection, algorithm. This process is technically complex. It has been largely unsuccessful because RF signal-level quantification circuitry cannot react fast enough to RF amplitude changes, and in the selection algorithm, high-speed switching between discontinuous signals causes damaging transients.

Fortunately, advances in performance along with reductions in the cost of integrated circuitry have enabled the development of a low-cost, highly efficient diversity reception controller that overcomes previous multi-antenna and multi-receiver problems. The controller's *ideal selection diversity* uses multiple receiver outputs that are selected by comparing RF carrier amplitudes.

Figure 2 above is a block diagram of the

diversity reception controller, which includes three main sections: an amplitude-sensing section, an analog logic section, and an output filter section.

The amplitude-sensing section contains high-speed, integrated, intermediate-frequency (IF) processors that provide IF amplitude information proportional to the received signal at each antenna. The processors generate an ultra-fast dc voltage received signal strength indicator (RSSI) proportional to the log of received power in decibels over a wide dynamic range.

The analog logic section processes the IF amplitude voltage, receives recovered modulation from the receivers and continuously selects recovered modulation from the receiver with the higher relative received signal.

The output filter section removes transients caused by high-speed switching between two discontinuous signals while ensuring constant input-to-output delays and fast settling characteristics. The filter can be programmed to accommodate all common IF bandwidths and response characteristics.

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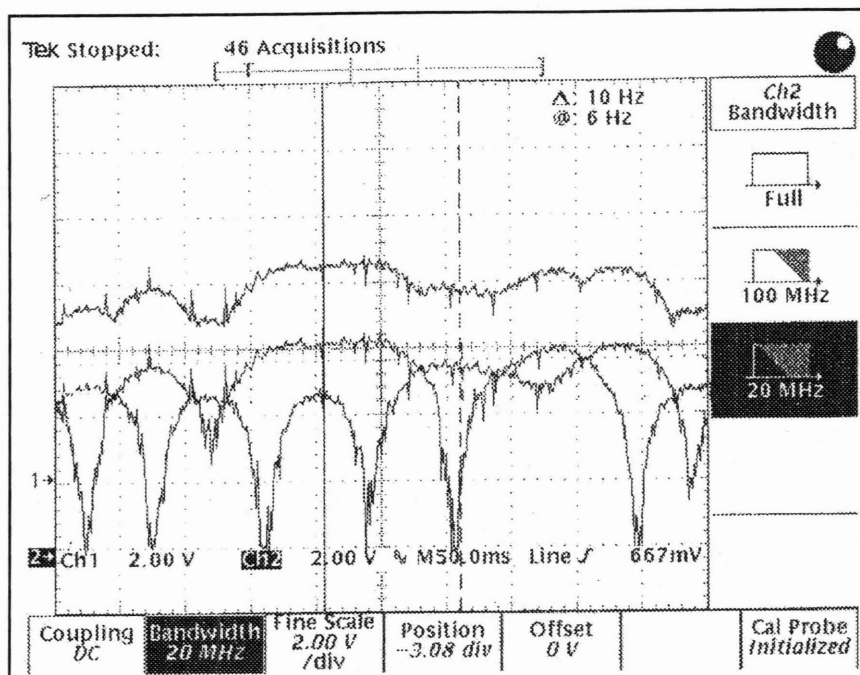


Photo 1. This digital storage oscilloscope display demonstrates the short-term fade reduction that an ideal selection diversity system provides during 500ms of reception in a vehicle traveling at 10mph. The lowest received signal strength encountered by the selected receivers was about -105dBm. Unselected receivers encountered numerous fades, six of which dipped to -130dBm.

Photo 1 at the left and Photo 2 on page 42 are plots taken from a digital storage oscilloscope connected to a diversity reception controller. These plots demonstrate short-term fade reduction that an ideal selection diversity system provides.

Ch1 (the upper trace) is the receivers' RSSI signal selected by the diversity reception controller to supply recovered modulation. Ch2 (the lower trace) displays the RSSI signals of the individual receivers. The diversity reception controller always selects the receiver with the higher relative signal level. This plot was taken in a vehicle traveling about 10mph. The plot is a 500ms snapshot of the short-term fading typically encountered in the mobile environment.

Voltages displayed on the oscilloscope plots correspond to the received signal amplitude as follows:

8.0V	-90dBm
6.0V	-100dBm
4.0V	-110dBm
2.0V	-120dBm
0.0V	-130dBm

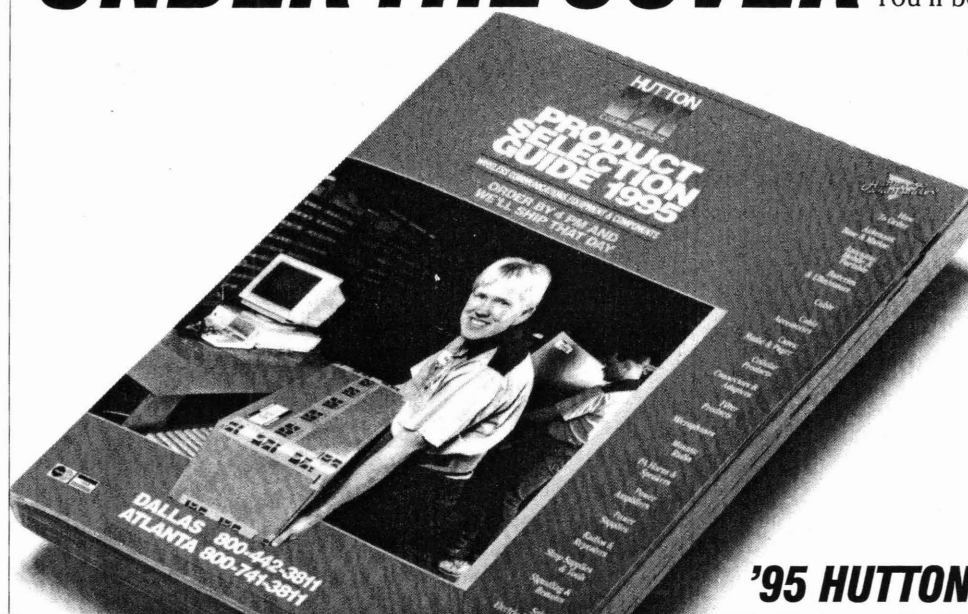
As shown in Photo 1, the lowest received signal strength encountered by the selected receivers was about -105dBm.

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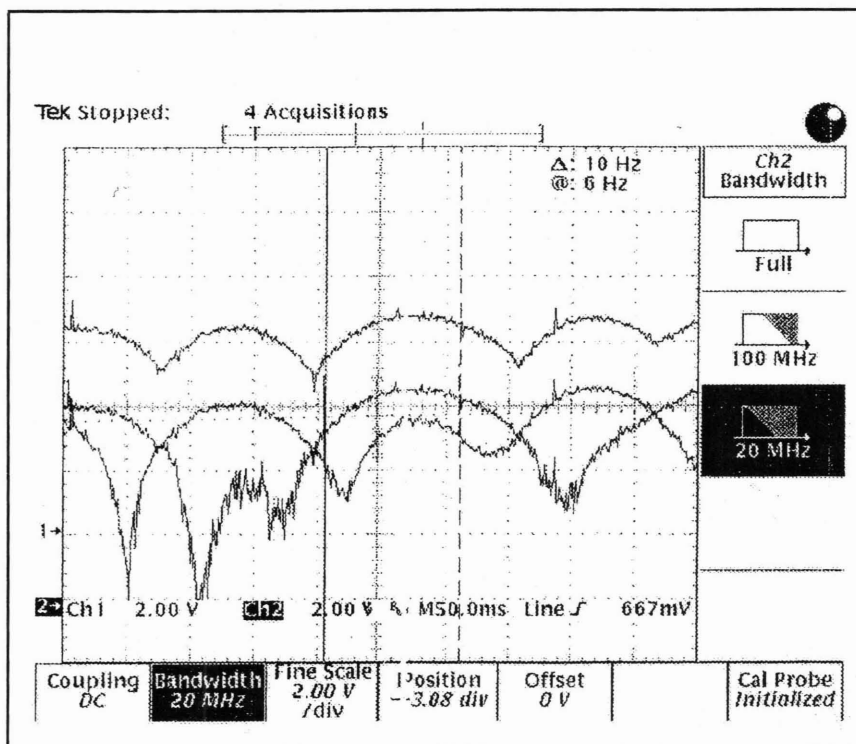


Photo 2. Another 500ms snapshot of typical short-term fading, taken at a slightly slower speed. Clearly evident is the classic shape of a Rayleigh distribution.

Unselected receivers encountered numerous fades, six of which dipped to -130dBm.

Photo 2 is another 500ms snapshot of typical short-term fading, taken at a slightly slower speed. Clearly evident is the classic shape of a Rayleigh distribution.

This diversity reception method's gain is proportional to the fade depth. For example, the percentage of probability of encountering a -10dB fade with ideal selection diversity reception is 1.0%, one-tenth of the percentage of probability without diversity. The percentage of probability of encountering a fade of -20dB with ideal selection diversity reception is 0.01%, one-hundredth of the percentage of probability without diversity.

Figure 3 on page 44 depicts this increase in fade protection. The left side of the graph indicates the percentage of probability of encountering a fade without diversity reception. The right side indicates the percentage of probability with a diversity reception controller.

The increase in performance provided by the diversity controller translates into system reliability. As seen in Figure 3, 99% reliability can be achieved with only a 10dB fade margin, representing a 10dB

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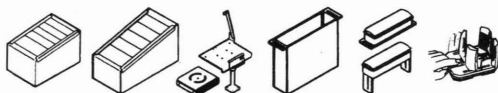
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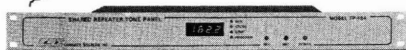
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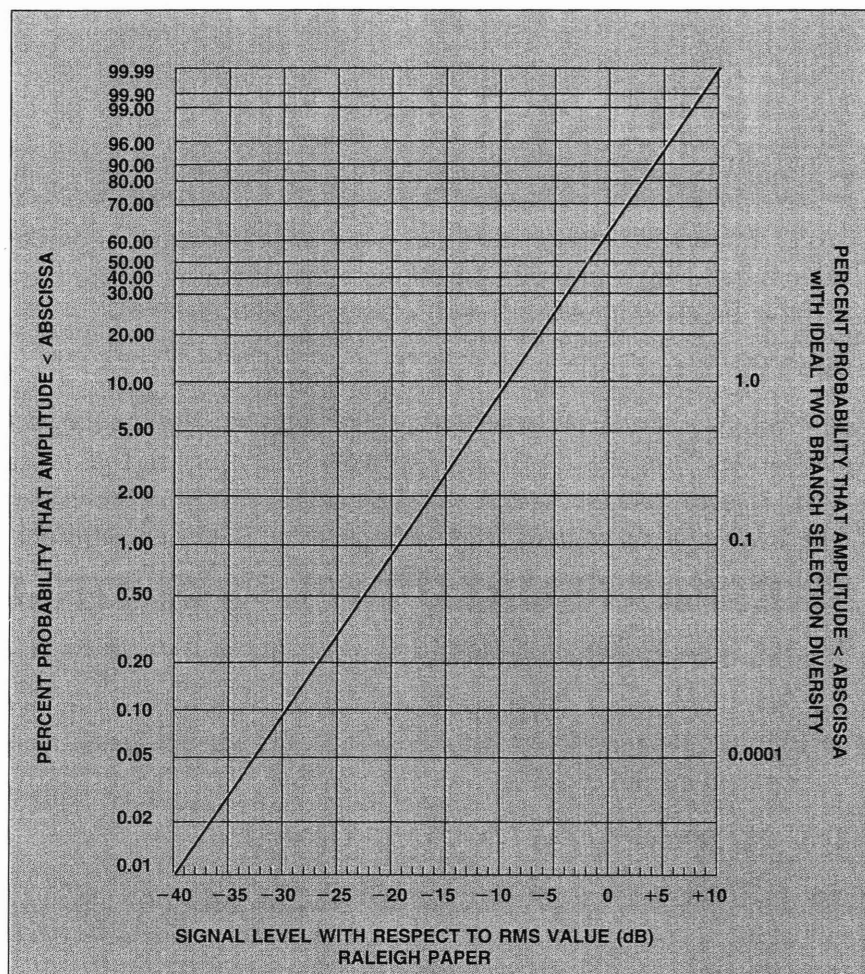


Figure 3. The increase in fade protection for ideal selection diversity reception is proportional to the depth of fades encountered. The left side of the graph indicates the percentage of probability of encountering a fade without diversity reception. The right side indicates the percentage of probability with a diversity reception controller.

improvement over a system without diversity reception. A 99.9% reliability can be achieved with a 20dB fade margin; this represents a 20dB improvement over a system without diversity reception.

Adding diversity reception to an existing radio communications system increases its coverage area and increases its reliability within any given area. For data communications systems, the success rate of first-attempt delivery of data messages rises dramatically, regardless of vehicle speed.

For voice communication, the diversity reception controller provides an 8dB-10dB improvement in signal-to-noise ratio (S/N).

The diversity reception controller can retrofit existing radio systems with the performance advantage of ideal selection diversity reception. An entire fleet can undergo retrofit, or only specific vehicles (such as vehicles that frequent poor coverage areas).

Although the focus of this article has been mainly on the mobile receiver, the

base receiver or repeater receiver is subject to the same short-term fading effects as the mobile unit. In terms of performance vs. cost, equipping the system base or repeater station with diversity reception yields the most performance per dollar because the advantages of diversity reception are shared by all users.

A dramatic increase in performance of mobile-to-mobile communications is possible when both repeater and mobile receivers are equipped with the diversity reception controller. The uplink's recovered audio experiences an 8dB-10dB increase in S/N performance. This better-quality audio is repeated on the downlink to a diversity-equipped mobile receiver with an 8dB-10dB increase in S/N performance. The aggregate S/N performance increase can be as high as 16dB to 20dB.

The controller is easy to install with few connections required and is compatible with all existing communications receivers.

Diversity reception controller helps radio system performance

Part 1—Short-term (multipath) fading can fluctuate received signals as much as 50dB. Without an adequate fade margin or other solutions, high-speed data communications can be adversely affected or entirely ruined.

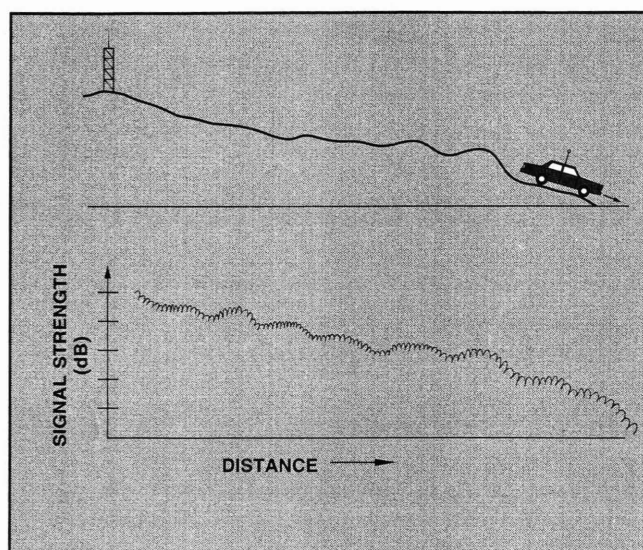


Figure 1. Long-term fading is caused primarily by distance and terrain. In general, as the distance between transmitter and receiver increases, the signal amplitude tends to decrease. Received signal levels tend to be stronger when the mobile receiver is on hilltops with a line-of-sight radio path to the transmitter. Signals are weaker when the line-of-sight path is blocked by terrain.

By Shane Fitzgerald

The best criterion for evaluating radio system performance is the mobile user's ability to communicate successfully, defined as meeting or exceeding a standard of performance.

A radio receiver's ability to meet a performance standard is a function of the received signal strength. If the received signal is at or greater than the level required, successful communication is possible. If the received signal is less than the re-

Fitzgerald is RF design engineer at ElectroCom Communications Systems, Santa Fe Springs, CA. ElectroCom manufactures the ideal selection diversity controller described in this two-part article.

quired level, successful communication is not possible.

Short of equipment failure, there is only one reason for the received signal amplitude to fall below the required level: *fading*. Fading is unquestionably the greatest cause of failed mobile radio communications. Reducing the effects of fading leads to significant improvements in radio system performance.

Two types of fading are present in the mobile radio environment, long-term fading and short-term fading.

Long-term fading is caused primarily by distance and terrain. In general, as the distance between transmitter and receiver increases, the signal amplitude tends to decrease. Received signal levels tend to be stronger when the mobile receiver is on hilltops with a line-of-sight radio path

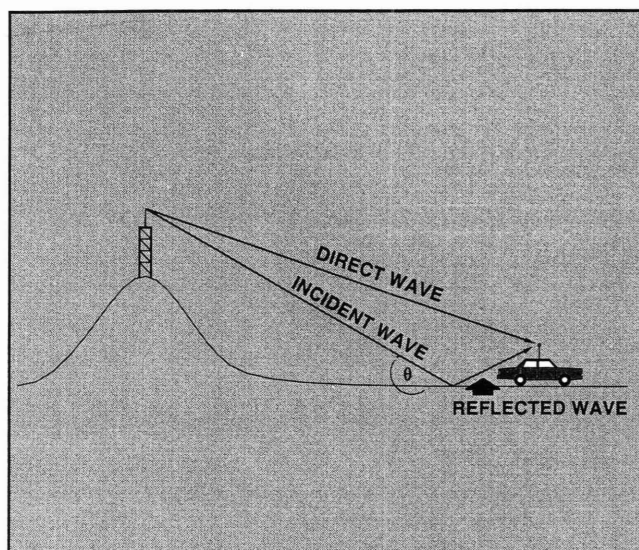


Figure 2. Short-term (multipath) fading is caused by a physical characteristic of electromagnetic waves. A direct wave travels from the transmit antenna to the mobile receiver antenna. In addition, an incident wave strikes the ground close to the vehicle and reflects upward and into the receiver antenna.

to the transmitter. Signals are weaker when the line-of-sight path is blocked by terrain. (See Figure 1 above left.)

Long-term fading has a gradual effect. As a receiver moves through a radio system's coverage area, the signal amplitude gradually moves up and down in response to distance and topography. The effects of long-term fading cannot be corrected at the mobile unit. Coverage deficiencies induced by long-term fading can be corrected only by changing the number or location of the radio sites, or by using on-frequency-repeater technology.

Short-term (multipath) fading is caused by a physical characteristic of electromagnetic waves. (See Figure 2 above right.) A direct wave travels from the transmit antenna to the mobile receiver antenna. In addition, an incident wave

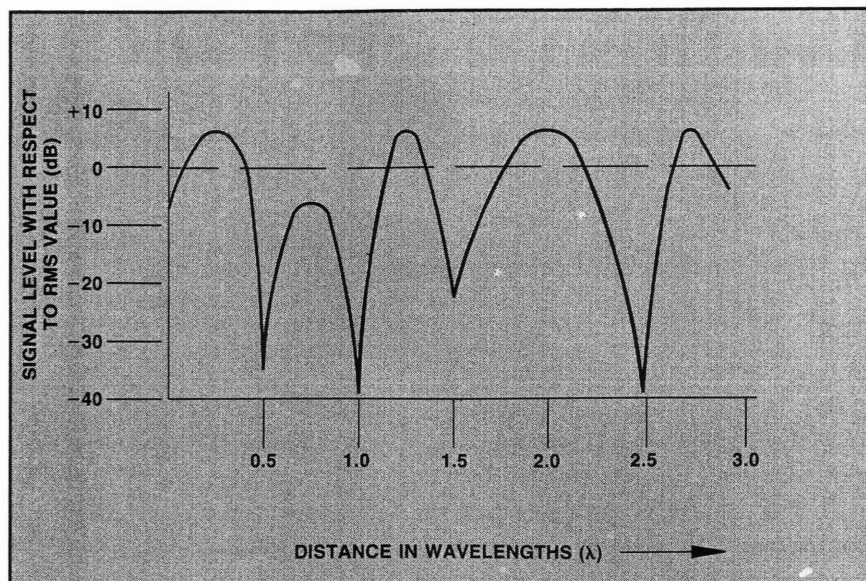


Figure 3. When a mobile receiver moves through an area, large fluctuations in received RF amplitude result. These fluctuations are periodic and occur at half-wavelength intervals (6.9 inches at 850MHz). The dynamic range of these fluctuations is about 50dB.

strikes the ground close to the vehicle and reflects upward and into the receiver antenna. The power in the reflected wave

depends on the angle of incidence (θ) and the ground's reflection coefficient. For average soil, the reflected wave's magni-

tude exceeds 0.9 when θ is less than 10° (which almost always is the case in land mobile communications). This small difference in magnitude means that the reflected wave's amplitude will be similar to the direct wave's amplitude. The mobile receiver responds to two sources of radio energy, the direct wave from a distance and the reflected wave from the ground in close proximity to the receiver.

When an electromagnetic wave strikes a reflector, its phase shifts 180° , a phenomenon known as specular or mirror-image reflection. These two waves, direct and reflected, interact to form a standing wave pattern of constructive and destructive interference.

Because it is impossible to see an electromagnetic standing wave pattern of constructive and destructive interference, it might be difficult to visualize. The water waves in a ripple tank demonstrate the phenomenon. (See Photo 1 on page 56).

In the ripple tank, two small floating balls (located at the center of the two circular patterns) are attached with strings to vertical oscillators on the bottom of the tank. As each ball oscillates up and down, it creates waves. Waves from the two

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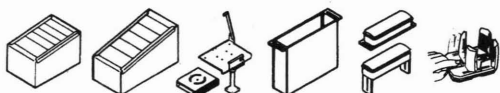
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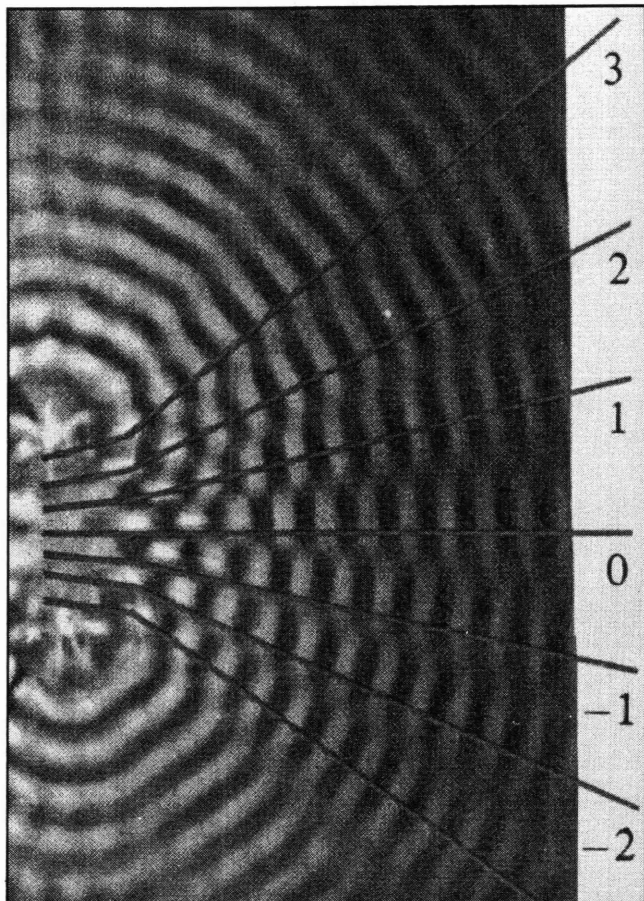


Photo 1. Because it is impossible to see an electromagnetic standing wave pattern of constructive and destructive interference, it might be difficult to visualize. The water waves in a ripple tank demonstrate the phenomenon.

sources react with each other to form the standing wave pattern visible in Photo 1.

This two-source interference pattern re-

sembles the direct and reflected radio-frequency (RF) standing-wave pattern created by two sources of radio energy. Clearly

visible in the photo are the lighter shades of the crests and the darker shades of the troughs of the standing-wave pattern.

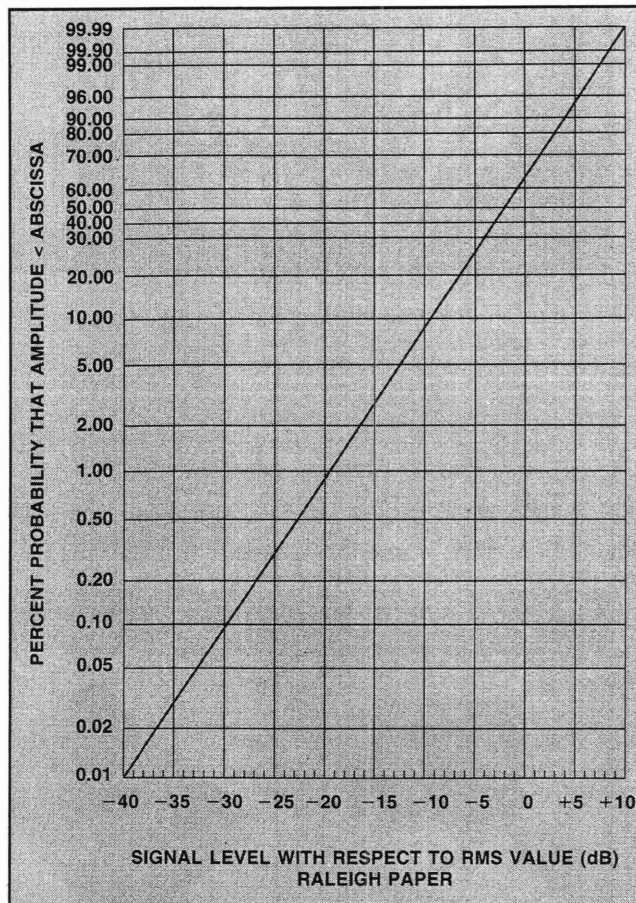


Figure 4. Signal amplitude at any given point ranges from 10dB above the average value to 40dB below the average value. This distribution is presented in its most useful form—Rayleigh paper.

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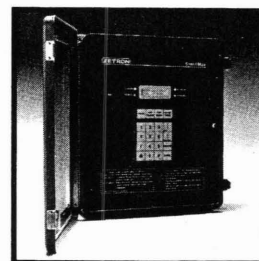
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The distance from maximum to minimum amplitude of the standing-wave pattern is one-half wavelength of the frequency of oscillation of the floating balls. Similarly, the distance from crest to crest or from trough to trough of short-term mobile radio fading is one-half wavelength of the radio waves' oscillation.

When a mobile receiver moves through an area, large fluctuations in received RF amplitude result. These fluctuations are periodic and occur at half-wavelength intervals (6.9 inches at 850MHz). The dynamic range of these fluctuations is about 50dB. Figure 3 on page 54 illustrates amplitude variations that a moving vehicle typically might encounter.

The distribution of short-term fading is a function of two random variables and is the sum of coherent sinusoidal waves of random phase and random amplitude. This sum describes exactly the mobile radio short-term fading phenomenon, which sometimes is referred to as *Rayleigh fading* after Lord John William Strutt Rayleigh (1842-1919), a British physicist who first described the distribution.

Signal amplitude at any given point ranges from 10dB above the average value to 40dB below the average value. This

distribution is presented in its most useful form—Rayleigh paper—in Figure 4 on page 56.

Rayleigh paper is an extremely useful tool for determining the percentage of probability that a signal will be above or below the average value. Numbers across the bottom of the graph represent signal amplitude (with respect to the average value) in decibels; therefore, 0 is the average value.

To find the percentage of probability of encountering a fade 20dB less than the average, follow the vertical -20dB line upward until it intersects with the diagonal line. Then read the graph directly to the left. In this example, there is a 1.0% probability of encountering a fade of 20dB.

Fade margin is an RF system design characteristic that indicates the system's ability to tolerate fades. It is expressed in decibels above the threshold where reliable communication takes place. As long as the average RF signal amplitude is 40dB above the level necessary for reliable communications—a figure that represents a 40dB *fade margin*—short-term fading will have no noticeable effect on the ability to communicate. Unfortunately, maintaining a 40dB fade margin in the field is practically impossible.

When a given geographic area's fade margin is 20dB, 1% of the locations within the area will have RF amplitude below what is required for communication. If the fade margin is 10dB, then 10% of the locations will have RF amplitudes below that required for communications.

For voice communication via radio, fade margin can be translated directly into system reliability because short-term fading effects on voice communication are more of a nuisance than a serious problem. The mind's processing power enables a listener to understand a voice message despite occasional noisy pops and hisses.

Short-term fading is extremely destructive to high-speed data communications. What the ear would endure as one little pop or a hissing sound may destroy several hundred bits of information. A mobile unit traveling at 15mph through a stationary environment will encounter 38 fades per second, and more if the environment includes moving objects such as trains, aircraft and automobiles. If these fades sink below the receiver's data recovery threshold, successful data reception becomes nearly impossible.

Next: How the integrity of data communications can be protected in a fading environment.

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Build a crystal filter to solve VHF interference

When adjacent-channel signals as close as 15kHz to the desired frequency interfere with reception, a crystal filter may offer the best solution to combine low insertion loss with high attenuation.

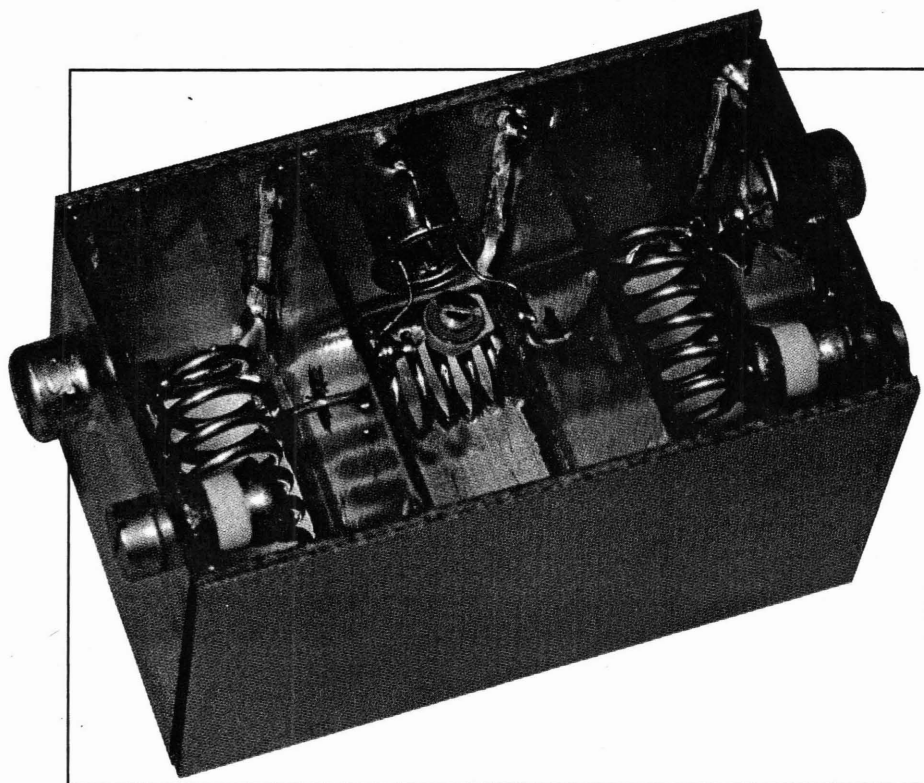
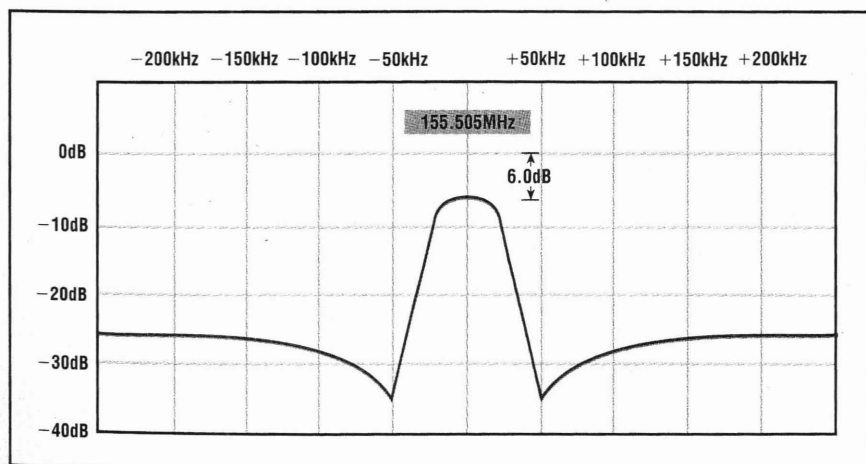


Photo 1. The single-filter, high-pass and low-pass filter of Figure 2 on page 56.



By Patrick E. Buller

One of the best solutions for adjacent-channel VHF highband interference is a crystal filter.

But where do you find one? There are very few crystal filter vendors listed in catalogs and product directories.

A homemade crystal filter can reduce by 25dB a VHF highband only 15kHz from the desired signal with only 1.0dB of insertion loss.

With more and more VHF stations appearing as if from nowhere, adjacent-channel interference has become a problem with no ready or reasonable solution. Large (10-inch) cavities configured as notch filters are inefficient for resolving interference caused by signals less than 100kHz from the desired channel frequency. A device with much higher "Q" is necessary for spacings closer than 100kHz.

High 'Q' solution

Crystals, with "Q" values higher than 70K, are the solution. A crystal filter can be configured either as a high-pass or low-pass notch filter with low insertion loss and, with a second crystal, as a symmetrical band-pass filter.

For many years, Motorola has offered

Figure 1. The Motorola model TLD6340A crystal selectivity curve with an 18.0kHz passband, 6.0dB insertion loss and 40dB maximum attenuation at a spacing of about 120kHz from the center frequency. Attenuation specifications are 13dB, adjacent channel, and 20dB minimum for alternate channels and greater.

Buller is an electronics design engineer for the Washington State Patrol, Bellevue, WA. He is a member of the Radio Club of America, IEEE, NARTE and APCO. He earned an amateur radio license in 1950, a commercial radio license in 1959 and a B.S.E.E. in 1963.

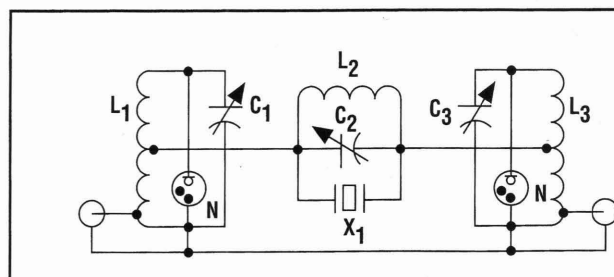


Figure 2A. This schematic diagram represents the best single-crystal, high-pass and low-pass filter based on the Washington State Patrol filter experiments. It uses an AT cut, 5th overtone crystal on 150MHz-160MHz frequencies.

SINGLE-CRYSTAL HIGH-PASS/LOW-PASS NOTCH FILTER

- L₁, L₃ 7 turns No. 16 AWG, 5/16" I.D., spaced 7/8", tapped at 3/4 turn input
- Output tapped at 3 1/2 turns from cold end for crystal
- L₂ 6 turns No. 18 AWG, 5/16" I.D., spaced 5/16"
- C₁, C₃ 10pF variable capacitor, Voltronics AP10SD
- C₂ 8pF ceramic trim capacitor
- N NE-2 neon bulb

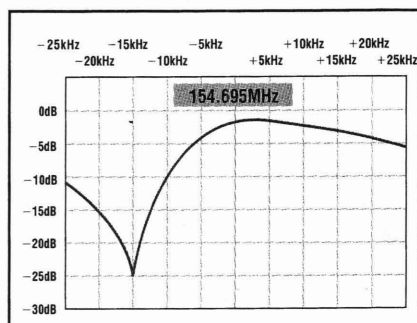


Figure 2B. The response curve of the filter adjusted for high-pass applications.

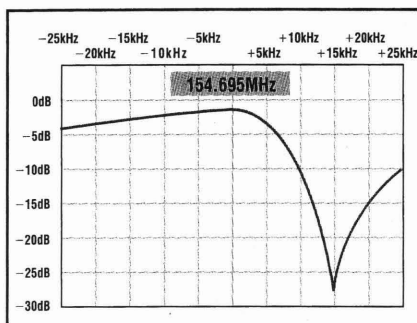


Figure 2C. The response curve of the filter adjusted for low-pass applications.

VHF crystal filters with a passband of 18.0kHz, 6.0dB of insertion loss and 40dB maximum attenuation at a spacing of about 120kHz from the center frequency. Attenuation specifications are 13dB, adjacent channel, and 20dB minimum for alternate channels and greater. Figure 1 on page 54 gives the 1972 specifications.

Crystals in the Motorola device can operate directly across a parallel-tuned circuit; nevertheless, attenuation is about 10dB. This filter has little value as protection from today's adjacent channel interference.

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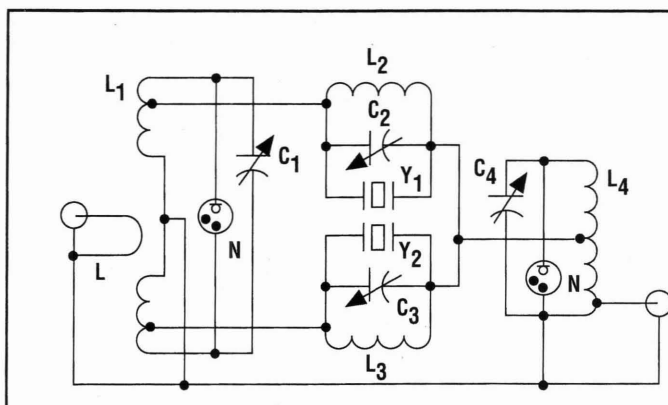
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- L₂, L₃ 6 turns No. 18 AWG, 5/16" I.D., spaced 5/8"
- L₄ 7 turns No. 16 AWG, 5/16" I.D., spaced 7/8"
- Input tapped at 3/4 turn
- Output tapped at 3 1/2 turns from cold end for crystal
- C₁, C₂, C₃ 10pF ceramic trimmer capacitor
- C₄ 10pF variable capacitor, Voltronics AP10SD
- N NE-2 neon bulb

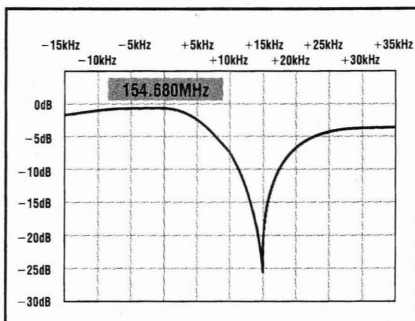


Figure 3B. The response curve adjusted for the high-attenuation, low-pass band-reject mode.

Figure 3A. The schematic diagram represents a half-lattice filter where the notch frequency is that of the two crystals. For greater bandwidth at the notch, one crystal should differ from the other by the required separation, perhaps 5kHz. The distinct advantage of this method is the quantity of notch attenuation obtainable—nearly 65dB with 2.0dB insertion loss, as shown in Figure 3B. Parallel resonant circuits L₂ and L₃ shift the crystal's resonant properties for the greatest attenuation.

Piezo Crystal, a vendor that was interested in cooperating in the Washington State Patrol filter experiments, can supply AT cut, 5th overtone crystals on 150MHz–160MHz frequencies. Many circuits were tried during the research and development stage that resulted in a more thorough understanding of crystal behavior in this ap-

plication. The best single-crystal, high-pass and low-pass filter is shown in Photo 1 on page 54 and Figure 2 on page 56. High-pass and low-pass characteristics are shown in Figures 2B and 2C.

Simplified design

The advantage of this design is that the

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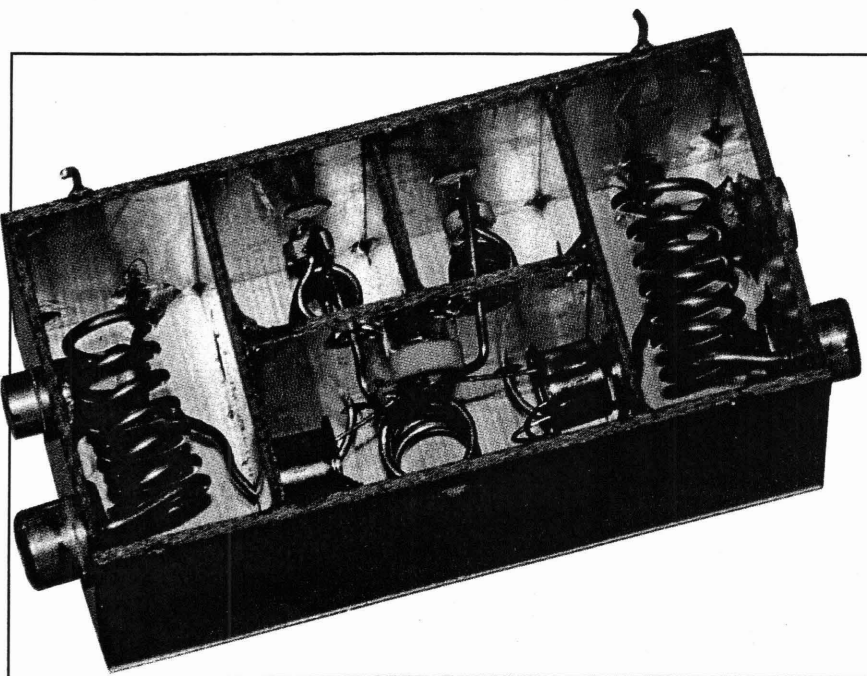


Photo 2. This picture shows the band-pass filter of Figure 4.

crystal determines the pass frequency, and its shunt element determines the notch frequency. This design resembles tunable cavities, used for duplexing and combining

where cavity tuning determines the pass frequency, and link tuning sets the reject frequency. This approach to crystal filter tuning simplifies the design. Choose the

crystal for your pass channel frequency and tune to the interference.

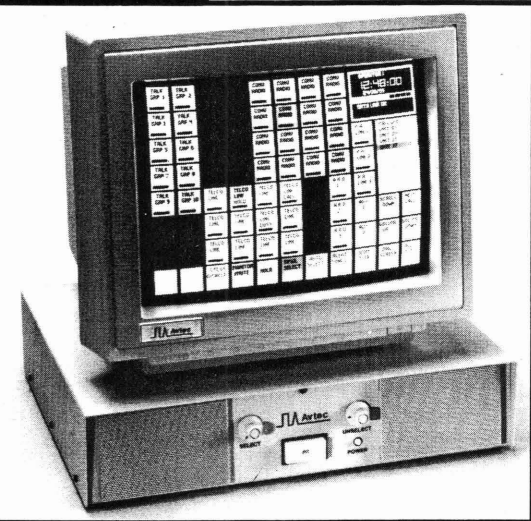
Input and output coils, L1 and L3 in Figure 2, match the 50Ω input and output line impedance to the Piezo crystal impedance, minimize the insertion loss and provide a small phase shift to compensate for slight crystal differences.

Tuning L2 from high-pass to low-pass, or the reverse, without tuning L1 and L3 does not yield low insertion loss. Changing the "tank Q" to fewer turns and increasing the capacity results in less flexibility in adjusting from high-pass to low-pass applications. All 20 crystals used for research and development matched the same impedance tap on the tuned coils.

Half-lattice filter

Figure 3 on page 58 is the schematic of a half-lattice filter, where the notch frequency is that of the two crystals. For greater bandwidth at the notch, one crystal should differ from the other by the required separation, perhaps 5kHz. The distinct advantage of this method is the quantity of notch attenuation obtainable—nearly 65dB with 2.0dB insertion loss, as shown in Figure 3A. Parallel resonant circuits L2 and L3 shift the crystal's resonant properties

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Figure 4A. The schematic diagram of a band-pass crystal filter with two crystals on the same frequency. Each crystal shunt element is tuned to the first adjacent channel above and the first adjacent channel below. The advantage of this circuit is that it can be tuned as a band-pass filter (Figure 4B) or as a band-reject filter (Figure 4C) with higher attenuation than what can be obtained with the single-crystal method.

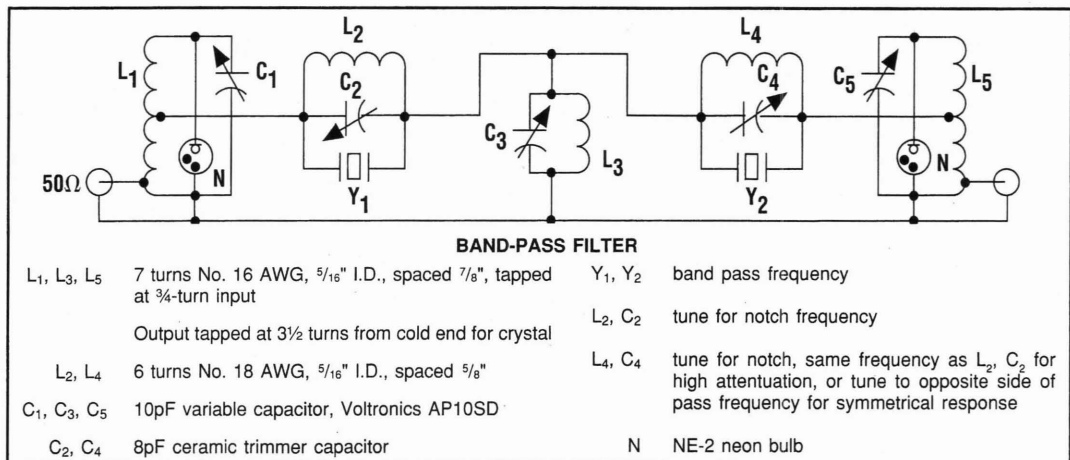


Figure 4B. Response curve of the two-crystal filter adjusted for band-pass applications.

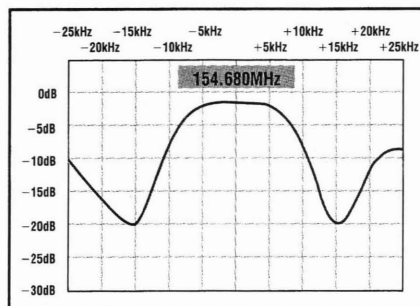
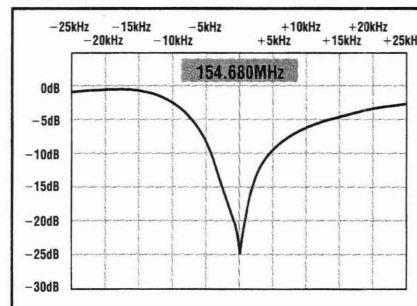


Figure 4C. Response curve of the two-crystal filter adjusted for a single-frequency notch.



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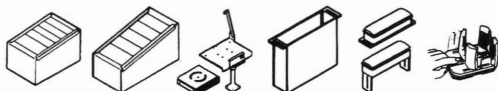
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for the greatest attenuation.

Photo 2 on page 60 and Figure 4 on page 62 represent a band-pass crystal filter with two crystals on the same frequency. Each crystal shunt element is tuned to the first adjacent channel above and the first adjacent channel below. The advantage of this circuit is that it can be tuned as a band-pass (Figure 4B) or as a band-reject filter with higher attenuation than the single-crystal method.

Interaction in tuned circuits L2 and L4 prevents uniform attenuation curves in the reject mode. (See Figure 4C.) Its attenuation is better than two filters in series (as shown in Figure 2) by about 0.75dB.

Filter characteristics

With a spectrum analyzer and a tracking generator, it is possible to display the filter's characteristics. Notice in Figure 4A that there are "spurs" (spurious responses)

above the crystal frequency at a lower level. These spurs have no effect upon the receiver's "on-frequency" performance.

Field tests with a 25dB notch filter, like the one in Figure 2, reduced a 15kHz adjacent signal from 300mV to 15mV, including the filter's 0.8dB insertion loss. The tests also proved that it interfaced in series with 10-inch cavities without additional loss.

The design of the Figure 2 filter was optimized for close-channel problems, although it can be tuned from the crystal frequency to approximately 100kHz on the low-frequency side and nearly 65kHz on the high-frequency side. For interference control at frequencies beyond 150kHz from the center frequency, there are other, more efficient, circuit arrangements.

Making adjustments

A spectrum analyzer with a tracking generator (or with signal generator and a tracking source) makes adjustments simple. If you are without these instruments, try this approach: Align input and output coils L1 and L2 of the filter, as in Figure 2, for maximum output signal at the received (crystal) frequency. Apply a signal at the desired notch frequency, and tune L3 for maximum attenuation at the output. Always have a valid 50V resistive load on each side of the filter for optimum performance.

Manufacturing 5th overtone crystals at this frequency is rather difficult. The first devices that were tried had little attenuation. At Piezo Crystal, it was theorized that the leads attached directly to the crystal element were too large, restricting performance.

The second-generation devices proved the theory. The 7th overtone has less notch attenuation than the 5th and 3rd overtones.

The enclosure for the crystal filters is made from double-sided printed circuit board stock soldered within each compartment. RCA phono jacks are used for in/out connections. Coils for the single-crystal filter shown in Figure 2 were wound on a common round pencil with a 5/16-inch diameter.

These filter ideas can be further developed to suit other needs. Someone probably will substitute toroids for the inductors, or perhaps slug-tuned coils. Use what you have.



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Radio path profiles and the Fresnel zone

By Harold Kinley, C.E.T.

Establishing a highly reliable radio communications link between two fixed points often requires a line-of-sight path between the transmitter and receiver antenna, especially if the transmitter power is to be held to a minimum. For radio propagation purposes, the radius of the earth is considered to be $\frac{4}{3}$ (1.333) times the natural radius. This longer radius produces a model of a "flatter" earth. (See Figure 1 to the right.) Thus, radio line-of-sight is approximately 15% greater than optical line-of-sight over smooth earth.

Formula 2 in the box on page 56 provides a means of calculating the maximum radio line-of-sight distance between two

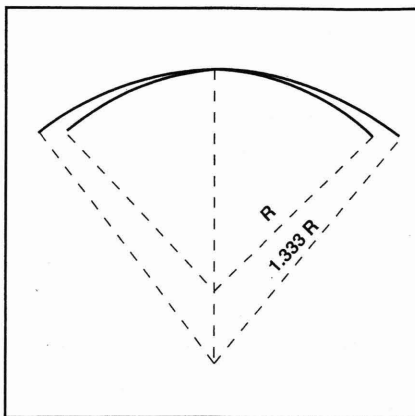


Figure 1. This graph shows how increasing the effective radius of the earth produces a 'flatter' earth, resulting in a longer line-of-sight distance between two points of given distance separation and antenna heights.

towers of given heights, h_1 and h_2 . Suppose the height of the first tower is 100 feet, and the height of the second tower is 150 feet. The maximum allowable radio line-of-sight distance (D) between the two towers is:

$$\begin{aligned} D &= \sqrt{2(100)} + \sqrt{2(150)} \\ &= \sqrt{200} + \sqrt{300} \\ &= 14.14 + 17.32 \\ &= 31.46 \text{ miles} \end{aligned}$$

continued on page 56

Kinley is a certified electronics technician with the South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985.

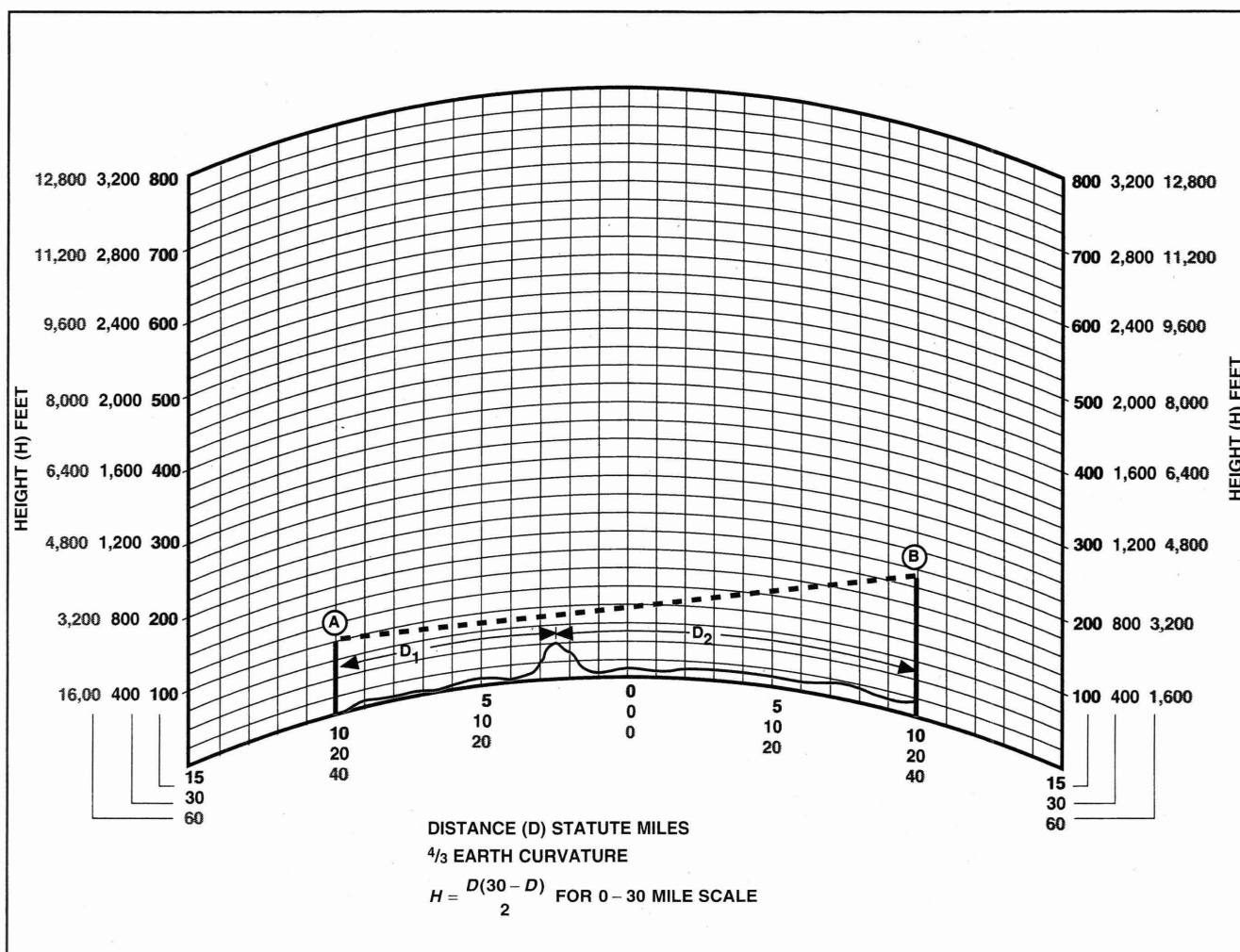


Figure 2. The $\frac{4}{3}$ earth curvature profile paper is used to check for proper line-of-sight clearance between antenna towers A and B. Although a direct line-of-sight path exists, the large hill probably would prevent suffi-

cient Fresnel zone clearance for free-space propagation loss. Profile paper courtesy of Ericsson GE Mobile Communications.

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continued from page 8

$$1) \quad D = \sqrt{2h}$$

where D = line of sight distance to horizon in miles
 h = tower height in feet

$$2) \quad D = \sqrt{2h_1} + \sqrt{2h_2}$$

where D = line of sight distance between two towers in miles
 h_1 = height of first tower in feet
 h_2 = height of second tower in feet

$$3) \quad r_1 = 31.67 \sqrt{\frac{\lambda D_1 D_2}{D_1 + D_2}}$$

where r_1 = first Fresnel zone in meters
 λ = wavelength in meters
 D_1 = shorter distance in kilometers
 D_2 = longer distance in kilometers

$$4) \quad r_1 = 548.33 \sqrt{\frac{D_1 D_2}{F(D_1 + D_2)}}$$

where r_1 = first Fresnel zone in meters
 F = frequency in MHz
 D_1 = shorter distance in kilometers
 D_2 = longer distance in kilometers

$$5) \quad r_1 = 2,280 \sqrt{\frac{D_1 D_2}{F(D_1 + D_2)}}$$

where r_1 = first Fresnel zone in meters
 F = frequency in MHz
 D_1 = shorter distance in miles
 D_2 = longer distance in miles

$$6) \quad L = 20 \log D + 20 \log F + 32.3$$

where L = free-space path loss in decibels between halfwave dipoles
 D = distance in miles
 F = frequency in MHz

$$7) \quad r_N = r_1 \sqrt{N}$$

where r_N = Nth Fresnel zone
 r_1 = first Fresnel zone
 N = integer (2, 3, ...)

Technically speaking

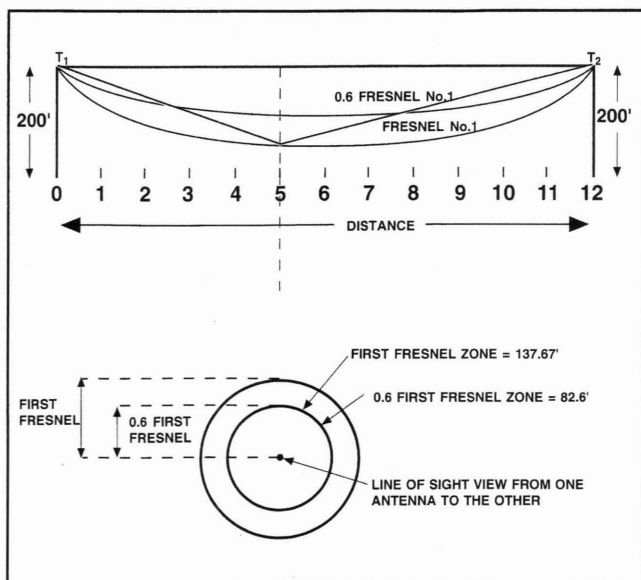


Figure 3. Towers T_1 and T_2 are separated by 12 miles. The curved lines represent the limits of the first Fresnel zone and 60% first Fresnel zone. The straight green line represents the direct path between the two antennas. The solid lines bounded by the first Fresnel zone represent the indirect path. The indirect path bounded by the first Fresnel zone is 0.5λ longer than the direct path. The first Fresnel zone completely surrounds the direct path as shown at the lower part of the figure. The actual shape of the Fresnel zone would resemble an elongated football surrounding the line-of-sight path.

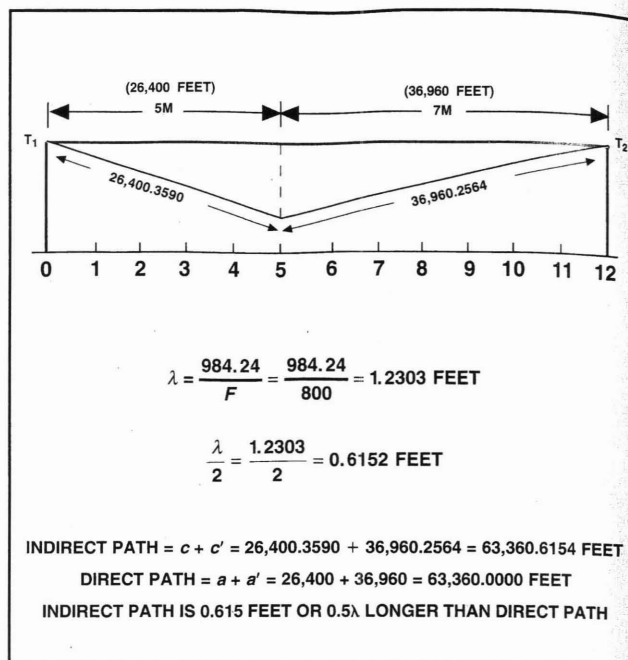


Figure 4. This drawing represents the direct and indirect path of the same towers shown in Figure 3. Line b , extending down from the direct path, represents the radius of the first Fresnel zone at a distance of 5 miles from tower T_1 . The direct path is equal to $a + a'$. The indirect path is equal to $c + c'$, and is 0.5λ longer than the direct path.

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Formula 2 in the sidebar can be rearranged to find the required height of the second tower if the required radio line-of-sight distance D and the height of the first tower (h_1) are known. It is rearranged as follows:

$$h_2 = \frac{(D - \sqrt{2h_1})^2}{2}$$

Remember, these calculations are for radio line-of-sight over smooth earth. Unfortunately, the earth is not very smooth in many places. The profile of the earth's terrain must be taken into account when establishing radio line-of-sight communication between two fixed points.

One way of constructing a path profile is to use a topographic map to compute the various elevations between the two fixed points and then plot them on $4/3$ earth curvature profile paper. (See Figure 2 on page 8.) Here, using the lowest scales, the height of tower A is 100 feet, and the height of tower B is 200 feet. The terrain profile (compiled from topographic maps) has been sketched in. Although there is a hill about 7.4 miles from tower A, there is a clear line-of-sight path between towers A and B.

Free-space attenuation

To achieve free-space attenuation (or

Technically speaking

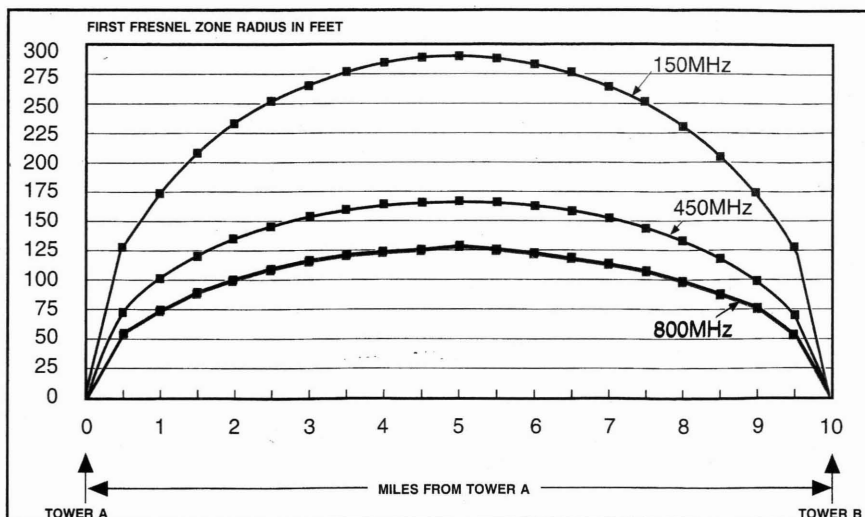


Figure 5. This graph shows the radius of the first Fresnel zone at three frequencies at any point between two towers separated by 10 miles. Notice how much smaller the first Fresnel zone radius is at 800MHz compared to 150MHz.

something close to it), an extra margin of clearance (more than radio line-of-sight) must be provided. For example, in Figure 2, even though radio line-of-sight exists between towers A and B, free-space attenuation is not achieved because of *diffraction* loss caused by the hill at distance

D_1 from tower A. To achieve almost free-space attenuation between towers A and B, at least 60% of the first Fresnel zone must clear the obstruction (hill).

Fresnel zone

The Fresnel zone is best described by

example. Figure 3 on page 58 shows the first Fresnel zone and 0.6 first Fresnel zone for a line-of-sight path between two antennas separated by 12 miles at a frequency of 800MHz. In practice, it is desirable to have at least 0.6 first Fresnel zone clearance to achieve almost free-space attenuation between two antennas.

In Figure 3, the first Fresnel zone radius extends 137.67 feet outward from the line-of-sight path at the five-mile point. This zone would completely encircle the line-of-sight path as shown at the bottom of Figure 3. The maximum extension of the first Fresnel zone radius would occur at the midpoint in the transmission path (at the six-mile point in the example in Figure 3). The direct path is indicated by the straight dashed line between towers T_1 and T_2 . The indirect path is indicated by the two solid lines that meet at the first Fresnel zone radius. For the first Fresnel zone, the indirect path is one-half wavelength ($\lambda/2$) longer than the direct path. This is shown more clearly in Figure 4 on page 58.

Referring to Figure 4, line b is the radius of the first Fresnel zone. The indirect path ($c + c'$) is 63,360.6154 feet. The direct path ($a + a'$) is 63,360.0000 feet. Thus, the indirect path is 0.6154 feet longer than the direct path. This distance (0.6154 feet) is equal to $\lambda/2$ at 800MHz.

At any point between the two towers (T_1 and T_2), where b is the radius of the first Fresnel zone, the indirect path ($c + c'$) will always be $\lambda/2$ longer than the direct path ($a + a'$). Distances c and c' are found from the Pythagorean theorem:

$$c = \sqrt{a^2 + b^2}$$

Frequency versus Fresnel zone

As the frequency increases, the radius of the first Fresnel zone becomes smaller, thus drawing closer to the direct line-of-sight path and requiring less clearance between the direct line-of-sight path and any obstruction. This reduction is illustrated by Figure 5 above, which shows the radius of the first Fresnel zone at three frequencies (150MHz, 450MHz and 800MHz) between two antennas 10 miles apart. Notice the difference between the first Fresnel zone radius at 150MHz and 800MHz.

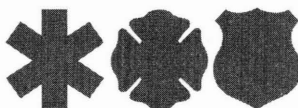
Computerized terrain profiles

As mentioned earlier, profile paper with $4/3$ earth curvature can be used to plot terrain profiles on a particular radial or azimuth between two fixed points. Even so, this requires careful interpretation of topographic maps, and many points must be plotted on the profile paper to give sufficient resolution of the profile. This can

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take up a lot of valuable time.

The alternative is to use a software package that includes terrain profile data (elevations). The software can generate a complete terrain profile graph in a short time while you take care of other business.

Figure 6 on page 64 shows a computer-generated profile path with a 0.6 first Fresnel zone radius added. The dashed lines in the figure were not generated by the program, but have been added to show

that 60% of the first Fresnel zone radius at a distance of 3.5 miles (from the antenna on the left) is 151.75 feet. This figure is the same as what would result from applying Formula 5 in the equations sidebar and then multiplying the result by 0.6 to get 60% of the first Fresnel zone.

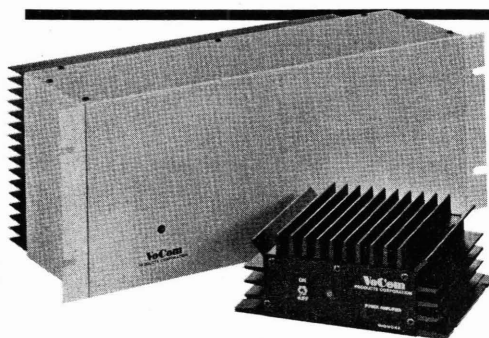
Figure 6 shows that the antenna elevations do not offer clearance of 60% first Fresnel zone between them. Therefore, the path loss will be greater than free-space

path loss. If clearance of 60% first Fresnel zone is desirable, the antenna height can be increased. This change is shown in Figure 7 on page 64 where the antenna height is 1,000 feet above mean-sea-level as compared to 910 feet in Figure 6. The result is a good clearance of 60% first Fresnel zone, which will result in minimal diffraction loss from the obstruction and which will yield almost free-space path loss between the two points.

If the operating frequency were 800MHz instead of 160MHz, the clearance would be 60% first Fresnel zone without raising the antenna height. (See Figure 8 on page 64.) Notice that 60% first Fresnel zone in Figure 8 is much smaller in radius than for the 160MHz 60% first Fresnel zone shown in Figure 6.

Figure 9 on page 64 shows a 50-mile path with the $\frac{4}{3}$ earth curvature drawn by

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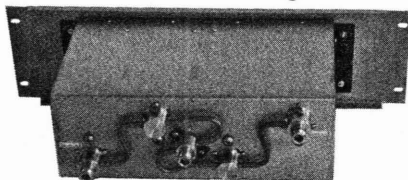
Even when line-of-sight paths are established with proper Fresnel zone clearance, problems can occur that yield results much worse than free-space propagation conditions.

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computer. Note the considerable mid-earth bulge at the midway point on the path. Even if the terrain were extremely smooth, considerable antenna heights would be required to overcome the mid-earth bulge.

Even when line-of-sight paths are established with proper Fresnel zone clearance, problems can occur that yield results much worse than free-space propagation conditions. Figure 10 on page 66 depicts a situation described to me by George Weimer of Trott Communications Group, Irving, TX. In the San Francisco area, under normal weather conditions a good, clear, line-of-sight path exists between the antennas in the example. During a temperature inversion, a thick fog layer forms and reflects radio waves (as shown by the dashed lines), resulting in a significant increase in the path loss between the two antennas.

The solution is to allow enough *fade margin* in the design to cover such anomalies. Too much built-in fade margin drives up the cost, whereas too little may severely reduce the path reliability, causing

Technically speaking

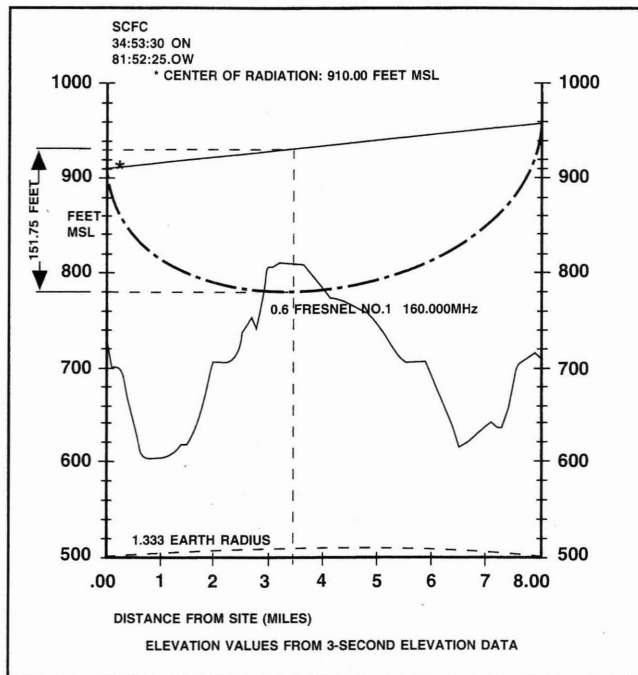


Figure 6. This terrain profile graph shows that a line-of-sight path exists between the tower on left and tower on right. Even so, at 160MHz, the path does not include 60% clearance of the first Fresnel zone. This graph was plotted with SoftWright's terrain analysis package (TAP) software.

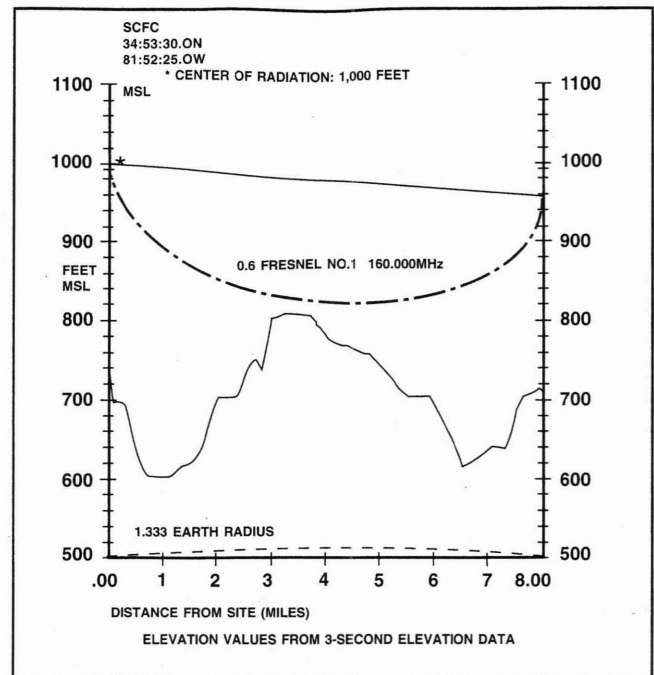


Figure 7. The tower on the left has been raised by 90 feet (compared to the same tower in Figure 6) to achieve more than 60% first Fresnel zone clearance over the obstruction 3.5 miles away. This graph was plotted with TAP software.

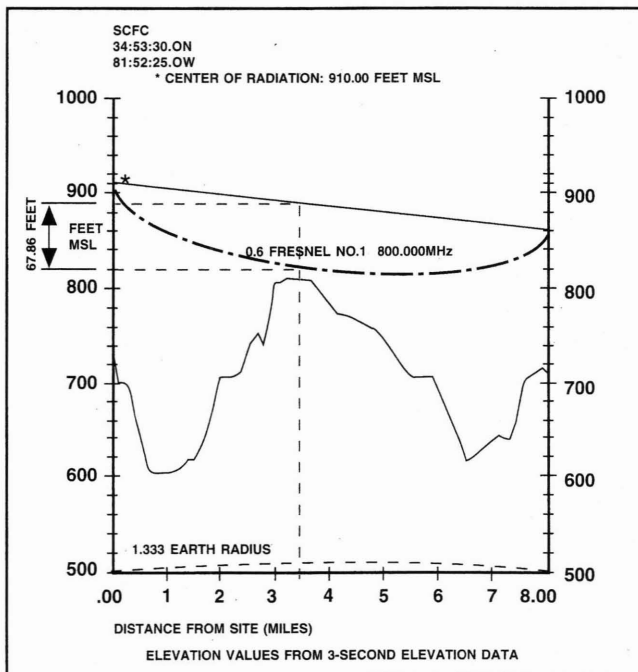


Figure 8. This terrain profile graph compares an 800MHz Fresnel zone and a 160MHz Fresnel zone. Note from Figure 7 that the left tower had to be 1,000 feet above mean sea level to achieve 60% first Fresnel zone clearance. Here, at 800MHz, 60% first Fresnel zone clearance is achieved with a tower height of 910 feet above mean sea level. The drawback is that the free-space path loss at 800MHz is about 14dB greater than the free-space path loss at 160MHz. This graph was plotted with TAP software.

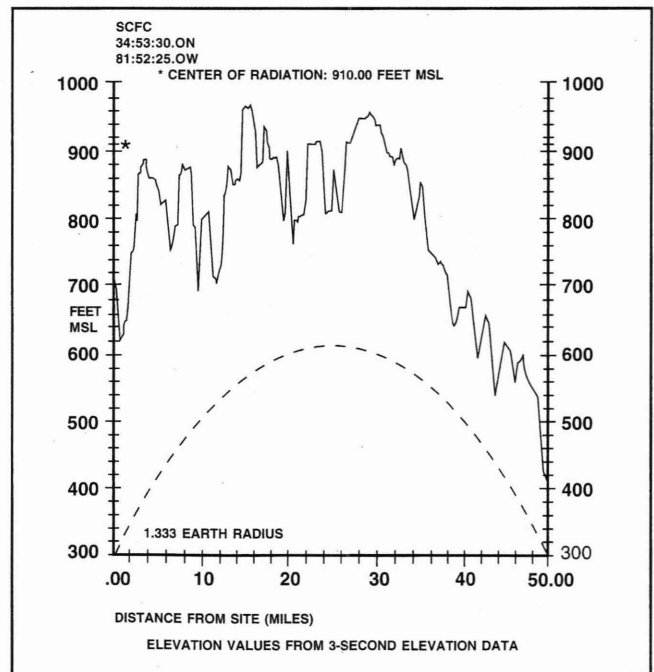


Figure 9. This terrain profile graph shows the tremendous effect of the earth's curvature over longer path distances. Tower heights greater than 300 feet would be required at each end of the path merely to clear the 'smooth earth's' curvature. Add to that the actual terrain profile, and the required tower heights for a line-of-sight path would be approximately 980 feet - 300 feet = 680 feet on each side. To achieve 60% first Fresnel zone clearance would require substantially greater tower heights. This graph was plotted with TAP software.

Technically speaking

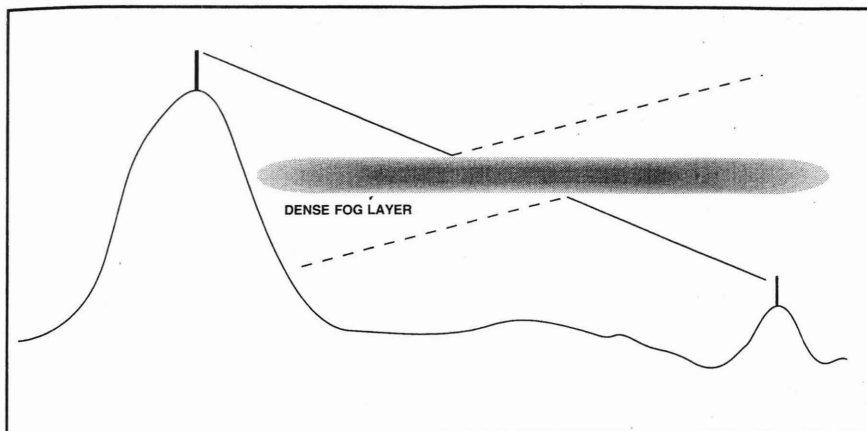


Figure 10. A dense fog layer has formed at a height between the two antenna heights. As a result, the signal from the higher antenna is reflected from the top of the layer, and the signal from the lower antenna is reflected from the bottom of the layer, resulting in significant attenuation to the path.

temporary interruptions in communications. Fade margins of 25dB to 30dB are often adequate, but there are special circumstances (such as this fog example) where a much higher fade margin might be required if the communications path is to be highly reliable. Use a common-sense approach along with your technical knowledge and data to find the "happy medium"

level between overdesign and underdesign.

For those who would like computer programs that compute radio-horizon, first Fresnel zone and free-space path loss, I just happen to have some available! These programs compute the radio line-of-sight distance between two towers of specified height (smooth earth) or between a single tower of specified height and the horizon.

The Fresnel programs compute the first Fresnel zone and 60% first Fresnel zone using meters, kilometers, feet, miles, wavelength and frequency. You can choose the measurement units from the menu. The free-space path loss program computes free-space path loss between two halfwave dipole antennas. Two additional programs are included to make conversions between wavelength (meters and feet) and frequency in megahertz as well as conversions between kilometers, miles, meters and feet—five programs in all.

These programs are for IBM-compatible computers running MS-DOS and are available on a single floppy diskette. Write to the author at 204 Tanglewylde Drive, Spartanburg, SC 29301-2949. Specify 5¼" or 3½" floppy. The cost is \$10 for either size diskette plus \$2.50 for shipping and handling.

Special thanks to Mike Wiebe of SoftWright for providing assistance with the terrain analysis profiles and to George Weimer of Trott Communications Group for answering specific questions. *Stay tuned!*



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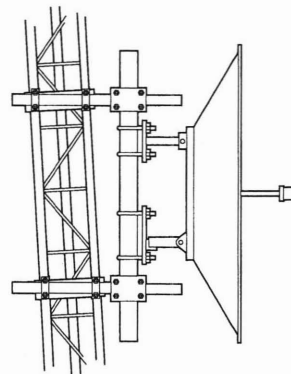
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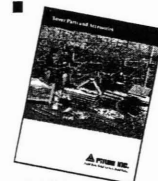
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How to use duplexers: Fine-tuning adjustments

Part 4—After initially tuning a filter using a spectrum analyzer and the technique described in Part 3, follow these fine-tuning steps to finish the job. Inexpensive test equipment can be used.

By Brian J. Henderson, P. Eng.

The best method for tuning filters uses thousands of dollars worth of test equipment.

Another, less expensive method, uses a transmitter, receiver, directional wattmeter and signal generator. One or more dummy loads may be required for a duplexer's unused ports.

Several rules and facts apply to this tuning method.

(1) All duplexer ports, whether or not they are connected to a signal generator, transmitter or receiver, must be terminated in a 50Ω impedance.

This rule cannot be stressed enough. Failure to follow this one simple rule results in poor filter tuning and poor system performance.

(2) When tuning any bandpass cavity, the lowest SWR point occurs at the passband frequency.

(3) Once the tuning rod is adjusted for the correct frequency in a pass or reject cavity, capacitor adjustment does not affect this tuning.

That is, once the tuning rod is set for pass frequency, this adjustment is not affected further by adjustment of the tuning capacitor, other ports of the duplexer string or other cavities.

Bandpass Duplexer

The first type of duplexer to be tuned is the bandpass duplexer. There is only one

adjustment per cavity. Connect transmitter, wattmeter and terminations as shown in Figure 1 below.

Set the transmitter to the correct pass frequency. Transmit into the duplexer string and adjust the tuning rods for minimum reflected power as indicated on the wattmeter. Note that reflected powers should be less than 1W, so it may be advisable to use a low-power wattmeter element.

Tune each bandpass cavity in the above manner. Then assemble all bandpass cavities together into the correct duplexer configuration. Fine-tune the transmit side as shown in Figure 2 below left.

Then reverse the transmitter and load connections. Set the transmitter to the receive frequency and adjust the receive side in the same manner. It is perfectly acceptable to transmit into the receive side of any duplexer, providing all terminations are in place and the receiver is not connected.

As a check for proper isolation, connect the signal generator, receiver and terminations as shown in Figure 3 below right. Set the signal generator and receiver to the same frequency.

Inject the signal into the duplexer antenna port. In this example, the signal

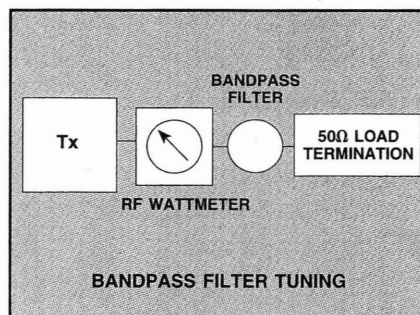


Figure 1. To configure equipment for tuning a bandpass duplexer, connect a transmitter, a wattmeter and terminations as shown.

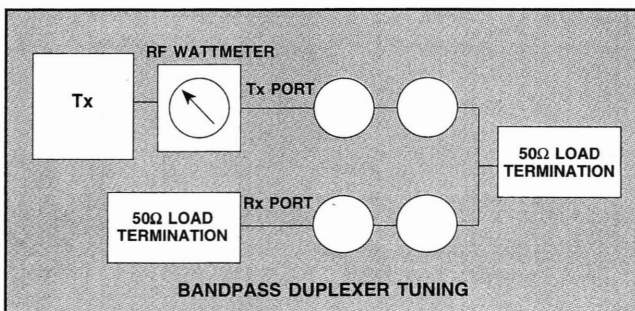


Figure 2. After tuning each bandpass cavity individually, assemble them into the correct duplexer configuration. Fine-tune the transmit side as shown. Then reverse the transmitter and load connections. Set the transmitter to the receive frequency and adjust the receive side in the same manner.

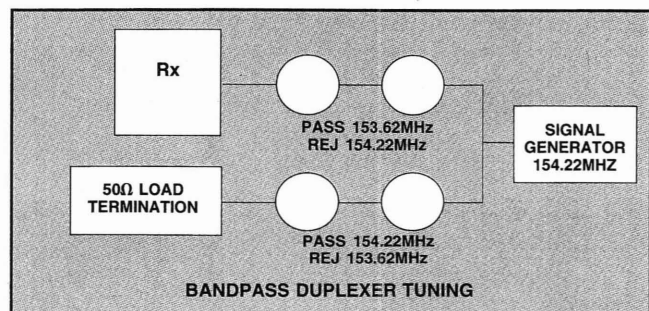


Figure 3. As a check for proper isolation, connect the signal generator, receiver and terminations as shown. Follow steps described in the text to measure the duplexer isolation.

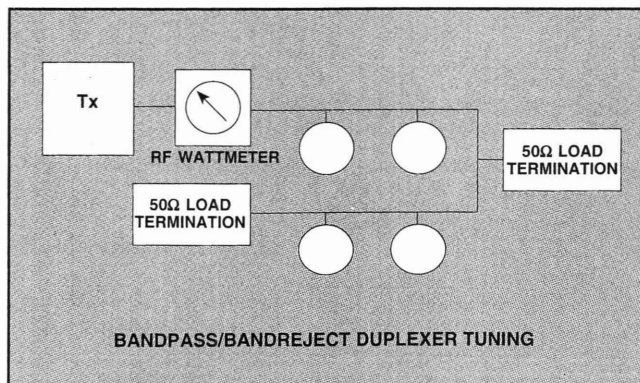


Figure 4. Steps for tuning the bandpass/bandreject duplexer are similar to those for the bandpass duplexer. Set up the duplexer as shown, and follow the steps described in the text.

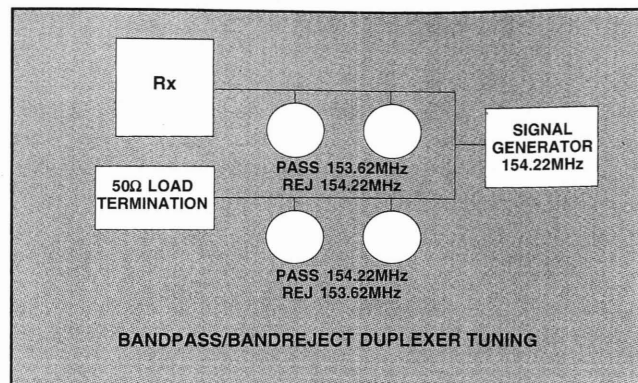


Figure 5. Use this equipment configuration to measure the receiver noise threshold in Step 2 of the bandpass/bandreject duplexer adjustment procedure.

generator and receiver are set to 153.62MHz. Determine receiver *noise threshold* (the point where noise begins to disappear). Record this reading in dBm.

Move the signal generator to the duplexer transmit port. Retune the signal generator and receiver to 154.22MHz. Increase generator output until the same noise threshold point is found. Again, record the generator output level in dBm.

The difference between these two levels

is the duplexer isolation in dB at that frequency.

Bandpass/bandreject duplexer

Tuning this type of duplexer is similar to the bandpass type.

Two adjustments are made on each cavity, the tuning rod and the capacitor adjustment. Set up the duplexer as shown in Figure 4 above left. In all cases, bandpass tuning rods are adjusted first; capacitors are

adjusted last.

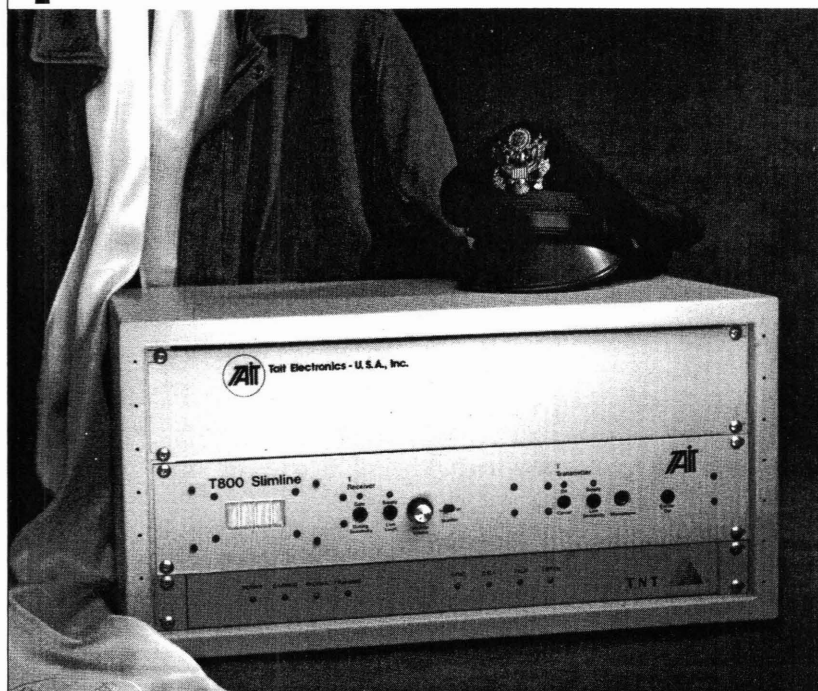
(1) Transmit into the duplexer on the correct frequency.

Adjust the tuning rods for minimum reflected power. Tune both "sides" of the duplexer—transmit and receive—on the correct frequency.

(2) Connect the signal generator to the antenna port. (See Figure 5 above right.)

Inject signal on the other frequency and adjust the capacitor for maximum rejection

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of this signal.

Open the receiver squelch.

Increase signal generator output until normal FM quieting is heard. Reduce generator output until noise returns. This is referred to as the receiver *noise threshold*.

(3) Adjust the capacitor for maximum noise as heard in the receiver. (This point represents the maximum attenuation of signal from the signal generator.)

(4) Keep increasing generator output and

adjusting capacitors until no further increase in rejection can be obtained.

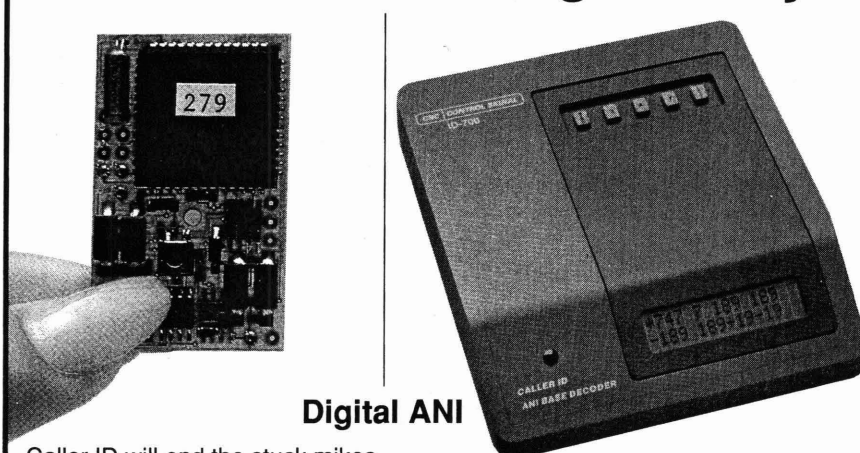
(5) Repeat this process for the other "half" of the duplexer string.

You may find one of these types of duplexers impossible to tune. If so, check the duplexer with a spectrum analyzer and tracking generator. The duplexer may have been "flipped" from a highpass to a lowpass or vice versa. See Figure 6 to the right for the proper appearance of highpass and

lowpass curves as shown on a spectrum analyzer display.

Remember that tuning a duplexer with a spectrum analyzer is *approximate*. Duplexers must be *fine-tuned* with the transmitter and receiver technique detailed above.

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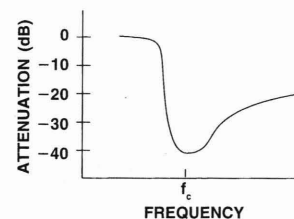
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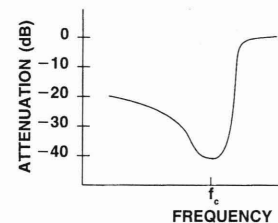
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(A) FREQUENCY RESPONSE
LOWPASS CAVITY



(B) FREQUENCY RESPONSE
HIGHPASS CAVITY

Figure 6. The spectrum analyzer display for a lowpass curve should look like curve (A), and the highpass curve should look like curve (B).

Next: Transmitter combiners and receiver multicouplers allow multiple repeaters to share the same antenna. If the system can tolerate the energy loss, combiners allow more efficient use of antenna space.

Duplexer series

Previous installments in this article series include:

- "How To Use Duplexers: Isolation Requirements" in the July 1994 issue.
- "How To Use Duplexers: The Various Types" in the August 1994 issue.
- "How To Use Duplexers: Installing and Tuning" in the September 1994 issue.

Back issues printed within the past two years can be ordered for \$5 each, postpaid. Call customer service at 800-441-0294. Copies of articles printed more than two years ago are unavailable from the publisher.

Use direction-finding to trace interference

Pick the direction-finding technology you like the best and learn to use it. When the time comes and you have an urgent need to track down a signal, it will be too late if you haven't had some practice.

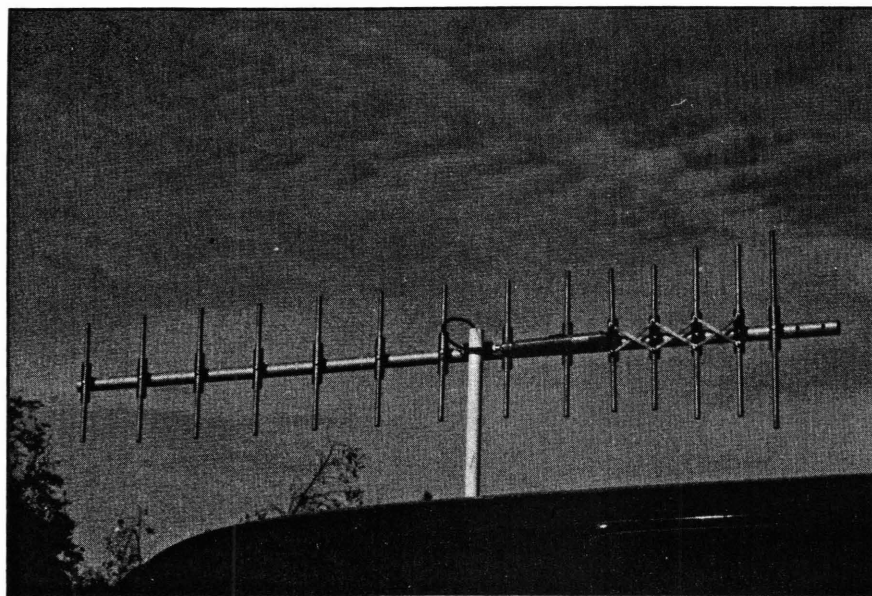


Photo 1. A simple, inexpensive way of taking a bearing to a signal from a mobile unit. This broadband yagi gives first-rate results from 450MHz to 470MHz. It rotates on a mast that protrudes from the vehicle win-

dow or through a hole in the roof. As the beam antenna is turned, usually by hand, a signal strength meter (S-meter) on the receiver indicates when the antenna points in the direction of the strongest signal.

By Joseph D. Moell, P.E.

There it is again! That strange carrier, interfering with your radio system. There is no modulation to identify it. Where is it coming from? How do you find the source?

Radio direction-finding (RDF) techniques used to trace the source of interfering signals were developed in the

early days of radio communications. Today, the military, the FCC and search and rescue agencies, such as the Civil Air Patrol and the U.S. Coast Guard, make heavy use of RDF.

Few radio shops are well equipped for efficient RDF. But you, as a radio technician or system operator, can put RDF to good use on your own behalf or on behalf of agencies and other customers you serve. Here are some examples:

- *Stuck transmitters*—Whether caused by equipment malfunction or by a user who sits on a hand mic, a continuous carrier on your frequency may halt all communication. The source must be

found—fast. If automatic number identification is not in use and no audible clues are heard, multiple-site RDF makes tracing the source much easier.

- *Stolen radios*—It is not uncommon for a lost or stolen hand-held or mobile radio to appear on the air, causing interference and harassment. An RDF network provides the best chance of rapidly stopping the interference and recovering the radio.

- *Surveillance*—RDF can locate criminals who use radios to coordinate their activities. Vehicles and sensitive cargo can be tracked on the move, without exposing officers to view. Tiny, remotely activated “bumper beeper” transmitters serve the purpose.

- *Lost individuals*—Mass media publicize the use of RDF to track emergency locator transmitters (ELTs) carried by aircraft and emergency position indicating radio beacons (EPIRBs) carried by boats. But sometimes, lives must be saved by tracking other transmitters.

Last year, when a single-engine airplane crashed in rugged terrain west of Denver, its avionics were destroyed and its ELT failed to activate. The pilot and passenger, both radio amateurs, used hand-held amateur VHF transceivers to call for help.

Three other amateur operators skilled in RDF quickly used their direction-finding equipment to determine the airplane's location. Evacuation of the two crash victims was complete within six hours.

- *Intermod products*—Although you may be able to listen to the offending intermod signal and identify the two-or-more mixing frequencies from the in-

Moell is head of the Transmitter Subsystems Section at Hughes Aircraft, Fullerton, CA. He is co-author of *Transmitter Hunting—Radio Direction Finding Simplified* (stock No. 2701); TAB Books, Blue Ridge Summit, PA 17294-0214. The 323-page handbook includes practical techniques and circuits for RDF.

formation it carries, knowing the frequencies does not necessarily pin down the source.

In Southern California, a strong mixing product interfered with a communications satellite downlink. The signal's components came from a non-commercial FM station and a VHF TV station's aural carrier. Both stations are on mountaintop sites. But RDF showed

that the mix occurred at an industrial area 25 miles from the mountain!

Calling for help

Why not just call the FCC for direction-finding service? The commission's field offices have sophisticated RDF equipment for many frequencies.

Unfortunately, the agency cannot guarantee a rapid response. Staff reduc-

tions and higher-priority requests for RDF capability, such as those involving critical government communications, may make the FCC's RDF capability unavailable to you when you need it.

It is far better to be prepared to hunt interference sources on your own.

You can simply wander around your coverage area, listening for a strong signal to indicate the transmitter you seek is close by. That is the so-called "hot-cold" method of transmitter hunting. A better name would be the "needle-in-a-haystack" method. It is not radio direction-finding, because there is no indication of direction. You need something that gives a bearing to the signal—a directional antenna—and you may have one already.

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- ☐ White House and presidential communications
- ☐ Communications involving safety of life and property
- ☐ Public safety and commercial communications
- ☐ Other

Yagi and quad antennas (beams)

Photo 1 on page 18 shows a simple, inexpensive way of taking a bearing to a signal from a mobile unit. Mount a yagi or cubical quad antenna on a rotating mast that protrudes from the vehicle window or through a hole in the roof. As the beam antenna is turned, usually by hand, a signal strength meter (S-meter) on the receiver indicates when the antenna points in the direction of the strongest signal.

Aluminum yagis are readily available and easily mounted. You may have extra yagis that can be pressed into service.

Quad antennas (See Photo 2 on page 22) can be made of stiff copper wire and PVC plastic pipe. Construction time is about two hours, plus time to modify the vehicle to attach the rotating mast and some sort of pointer and compass rose at the mast base.

The larger the beam, the better the bearing resolution. Usually four ele-

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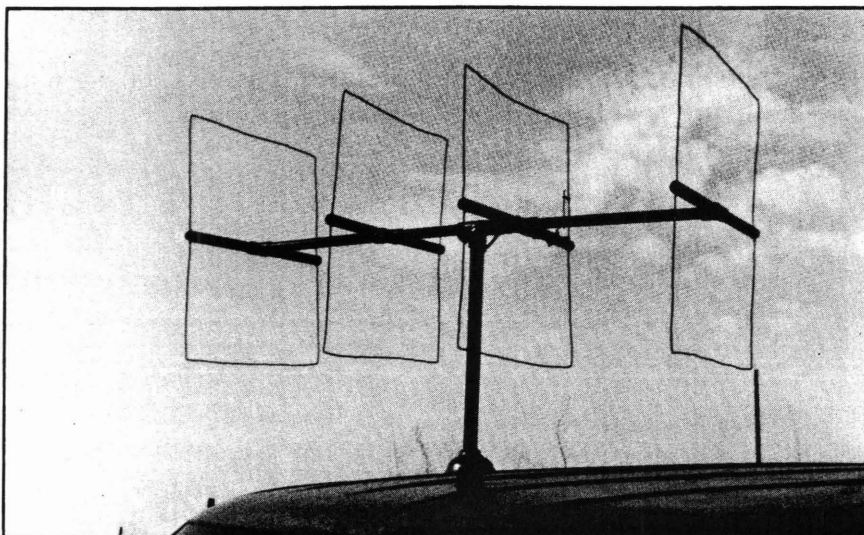


Photo 2. Cubical quad antennas can be made of stiff copper wire and PVC plastic pipe. Construction time is about two hours, plus time to modify the vehicle to attach the rotating mast and some sort of pointer and compass rose at the mast base.

ments on VHF highband is a reasonable number for suitable bearing accuracy and ease of turning. Frequency bandwidth for adequate directional performance is no greater than 10%. Thus, more

than one beam antenna may be required to cover all frequencies.

You need an S-meter on the receiver to determine accurately when the beam is pointing at the signal source. Exter-

nal S-meters can be added easily to receivers that do not have them.

Because FM receivers saturate in the presence of extremely strong signals, an external attenuator is a must. The attenuator reduces signal input to the receiver as you come closer and closer to the signal source.

Polarization

The orientation of the yagi elements and of the quad's feedpoint determines polarization. Your RDF beam must be polarized the same as the signal source to reduce adverse effects of multipath reflections.

Most land mobile transmitters use vertically polarized antennas. But intermod mixing points and transmitters in the broadcast and other services may have predominantly horizontal polarization. Make provisions for rotating the antenna boom to select between vertical and horizontal polarization as necessary.

The main advantage of a gain RDF antenna is its high sensitivity. When the signal is weak, a beam antenna outper-

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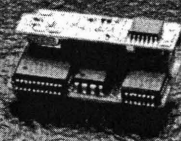
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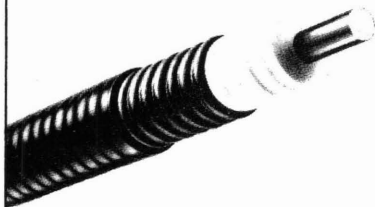
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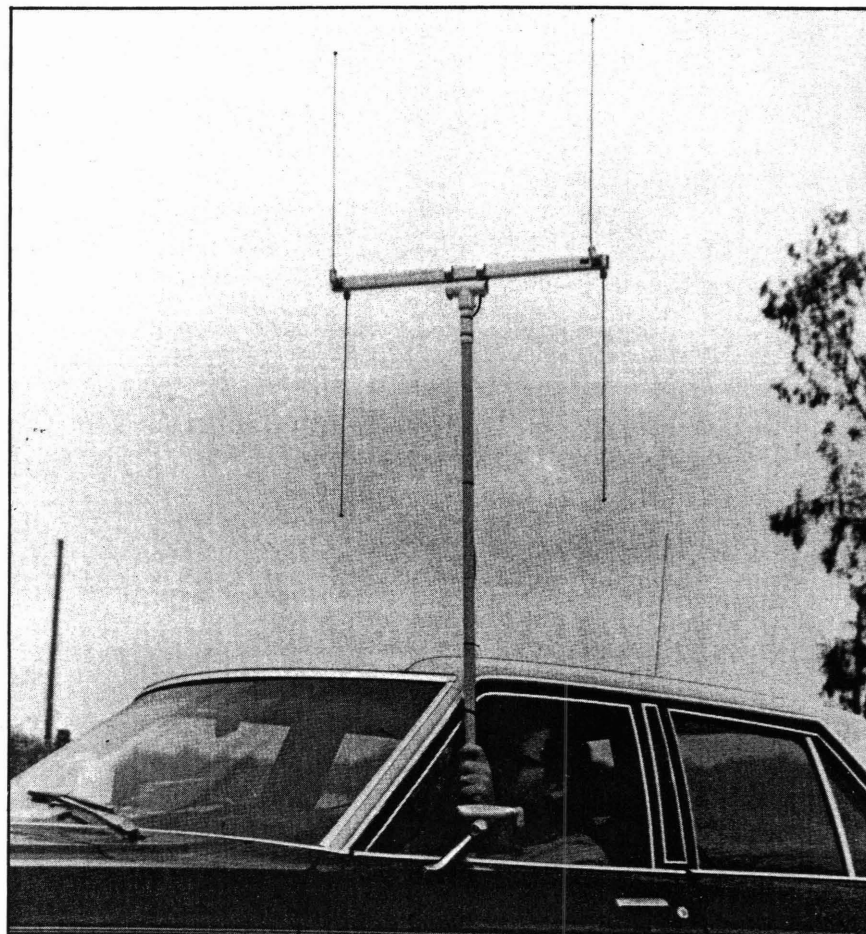


Photo 3. Homing RDF units use a pair of vertical antennas at the ends of a horizontal boom. The antennas are separated by less than a half-wavelength. The SuperDF made by BMG Engineering works on foot, or can be used this way on a vehicle.

forms other RDF methods. When you must find the source of a feeble spurious emission that just barely holds your hilltop receiver open, the beam antenna is your best bet.

Beams are unsuitable for use below 100MHz. Yagi elements for 33MHz VHF lowband are about 14 feet long. That length is not practical for the average squad car!

Other problems associated with beams are their requirement for manual turning. Manual turning, together with the usual signal flutter in mobile service, makes taking bearings a practiced art, particularly on intermittent signals. Having two individuals, one to drive and one to rotate the antenna, gives maximum efficiency.

Homing-type RDF sets

Homing RDF units use a pair of vertical antennas at the ends of a horizontal boom. The antennas are separated by

less than a half-wavelength (See Photo 3 above.) The antennas are switched back and forth electronically several hundred times per second. Two types of homing RDF sets operate on different principles, but they are used the same way.

Switched pattern units, such as the L-Per made by L-Tronics, use AM receivers and determine which of two antenna patterns produces a stronger signal.

Time difference of arrival (TDOA) units, such as the SuperDF made by BMG Electronics, work only with FM receivers. They use phase techniques to determine which of two antennas is closer to the signal source.

With both units, an indicator, such as a zero-center meter or a pair of light-emitting diodes (LEDs), tells the user which way to turn the antenna, left or right. A sharp "crossover" indication reveals when the antenna is pointed ex-

actly at the signal source.

For switched pattern units, crossover occurs when signals reach equal amplitude as received by each pattern. For TDOA units, crossover occurs when the antennas are equidistant from the source.

Both types are sold commercially at reasonable prices. The L-Per comes with a built-in, four-channel, crystal-

controlled receiver; an attenuator; and a meter. It is compact and comes ready to use, but you must have a crystal for the desired frequency. Frequencies can be selected in one of several bands from 120MHz to 320MHz. Prices start at about \$400.

The SuperDF connects to an FM receiver or scanner, so it works over a wide frequency range: 100MHz to

260MHz on the VHF antenna and 250MHz to 550MHz on the UHF antenna. It does not need any additional crystals.

The SuperDF can be carried on foot, together with a hand-held receiver. Alternatively, its rotatable antenna can be mounted on a vehicle for use with a mobile receiver. The assembled SuperDF unit with VHF antenna is less than \$300, not including a receiver.

The portability of homing RDF sets is advantageous. They are easy to set up and carry. When not in use, they can be folded up and stored in a small space in the vehicle.

Despite a lack of antenna gain, they have enough sensitivity.

Both are highly accurate in the absence of multipath. Inferior stationary multipath performance is the major disadvantage shared by these units. It is best to stay in motion to "average out" the bearing swings caused by signal reflections.

Homing RDF units sometimes give unreliable results when the signal is horizontally polarized. Special techniques must be used in such situations.

S-meters are not needed with homing RDF units, which are not swamped by strong signals. Usually, these characteristics are advantageous. But sometimes they turn into disadvantages. Because the units do not display amplitude information, the hunter may pass by the subject transmitter without noticing it. With the L-Per, overshooting the transmitter is less likely because its internal sensitivity control indicates proximity to the transmitter.

Airborne RDF

If aircraft are used in your operations, do not overlook the advantages of airborne RDF. Besides the ability to cover large areas rapidly without traffic or road problems, the use of a high platform eliminates most multipath problems that plague ground-level VHF RDF in uneven terrain. Airborne RDF really shines in search and rescue operations in rugged wilderness.

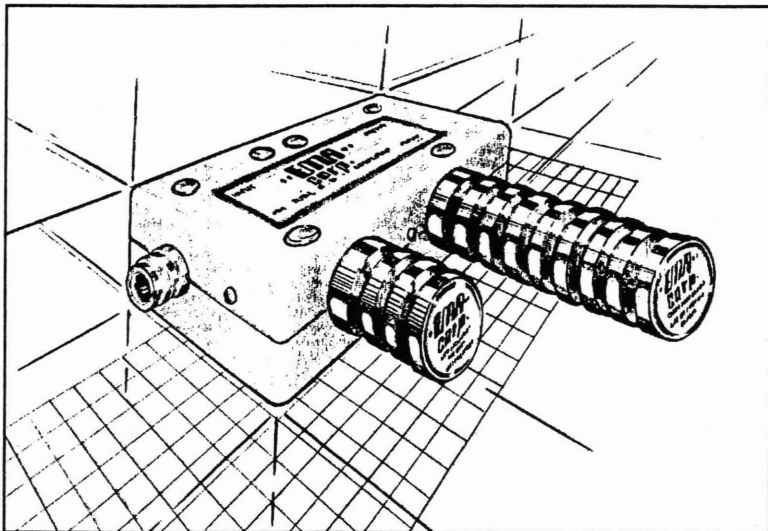
Homing RDF units, such as the L-Per or SuperDF, are ideal for airborne use because they indicate the direction to turn. Antennas can be mounted on the side of the cockpit or, less desirably, taped to the windshield.

Rotating antennas are unnecessary;

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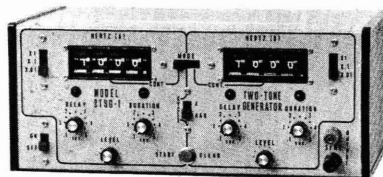
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just turn the plane! The pilot simply follows the RDF's left-right indications to fly right over the signal source. With practice, it is possible to maneuver a small plane to fly over the target with an accuracy of a few feet. Airborne RDF experts can determine from the air which car in a parking lot has the transmitter in it.

Ring antenna RDF sets

Just as there are two major types of homing RDF units, there are two types of ring antenna systems: Doppler and Adcock (sometimes called Watson-Watt). They use similar antennas and displays, but different processing techniques. One uses AM detection, the other, FM.

Ring antenna RDF units offer a major improvement in operating convenience compared to manually turned antennas. They are much easier to use, especially when hunting alone.

Ring antenna RDFs typically use a set of four or more identical vertical antennas. Dipoles are used for fixed installa-

tions. Quarterwave whips suffice for mobile use.

The mobile antenna system may be a single assembly fitted with suction cups to grip the vehicle roof, or four individual whips connected to a switcher

Direction-finding applications

- ☐ Recovering stolen radios
- ☐ Finding 'stuck mics'
- ☐ Tracking surveillance transmitters
- ☐ Search and rescue
- ☐ Tracing intermod sources

by coaxial cable lines of equal length. Individual magnetic-mount whips work fine. Either the suction-cup or magnetic-mount version is readily moved from one vehicle to another.

RDFs are popular for use with marine

VHF transceivers. Several companies make them, including Apelco, Koden and Simrad. Usually they are sold as add-ons for your own FM receiver, with no internal modifications necessary.

For the land-based two-way radio commercial market, Doppler Systems offers RDF sets with display and antenna for VHF highband beginning in the \$800 price class, not including the receiver.

RDFs using the Adcock principle require AM detection in the receiver. Often, the receiver is sold in a package with the RDF set, but an external receiver can be used, with minor modifications. Compudyne EWI Division supplies three-channel Adcock RDF sets under the name OAR for commercial use.

Both types of ring antenna RDF sets can be provided with a digital readout of bearing in degrees, or a less expensive circular display that "points" to the signal source direction. Doppler units typically use a ring of light-emitting diodes (LEDs). Adcock units generally



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feature a cathode-ray tube (CRT) display with "vectors" pointing in the direction of signal sources.

The CRT display is superior because it clearly indicates the presence of most types of multipath. Be prepared to pay a premium for the CRT display: Adcock RDF sets generally cost several times the price of Doppler units.

Doppler and Adcock RDF units are

favorites of many RDF technicians because of their rapid response. No mechanical turning is necessary, so the units display bearings even for short signal bursts. Compared to yagis and quads, Doppler and Adcock units are more broadband and are easier to install on almost any vehicle.

Some inexpensive Doppler units use PIN diodes to "hard-switch" among the

antennas. In areas with high RF levels, hard switching creates noise and cross-modulation products that reduce RDF sensitivity. Doppler Systems units use sophisticated "soft-switching" antenna selection circuits that minimize cross-modulation products and maximize sensitivity.

Ring antennas are not the answer for every RDF situation. They are not suitable for hand-carrying. They cannot be used to track pulsed or non-carrier signals, such as power line interference and single-sideband signals. Readout

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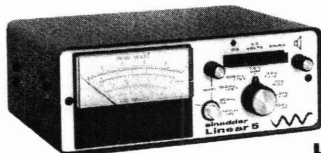
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Ring antennas are not the answer to every RDF situation.

resolution on some models is limited. Bearing error may be severe when the signal is horizontally polarized and when multipath is present. Sensitivity may be inadequate on weak signals.

Triangulation

Having read this far, you probably are convinced that equipping one of your vehicles with RDF gear could be important.

If your service area is small, one RDF vehicle might be all you need. But what if you service a large metropolitan system with satellite voting receivers, or a wide-area system with mountaintop repeaters? Your single vehicle is not likely to be in the right place at the right time to hear the source of the problem and respond rapidly.

When it is necessary to augment the single vehicle with additional RDF capability, it may be done by equipping more vehicles with RDF gear or by supplementing the mobile RDF unit with fixed RDF sites at strategic, elevated sites. Properly located base stations and vehicles can "triangulate" their bearings to reduce the search area dramatically.

Triangulation does not provide pinpoint accuracy. Bearing errors on VHF signals from elevated RDF sites can be

Direction-finding equipment

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significant, depending upon the equipment used and the local conditions.

For example, atmospheric inversions and the "coastal effect" in the Los Angeles basin cause VHF hilltop bearings over land and water paths to appear to drift 15° or more in a typical day.

With three RDF sets accurate to a bearing resolution of $\pm 10^\circ$ and separated equally from one another by 20 miles, bearings to a transmitter midway between them would define an area of

13.8 square miles. Thus, hilltop sites alone are not enough for a complete RDF system, but they provide a target area for RDF vehicles.

Get started now

Whatever RDF system you decide is best for your particular needs, now is the time to obtain it and learn to use it. Do not delay. When you have need for RDF, it will be too late.

No matter how simple your chosen

system seems to be, it will take some time to become familiar with its use. Terrain affects readings, so you should make trial runs to track down transmitters in known locations.

Once you are comfortable with the system operation, have someone transmit to you from an unknown location to test your RDF skills. Once you have had a few successful hunts, you will be ready when the need to track a signal to its source arises.

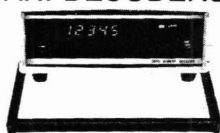
RDF as a sport can be lots of fun. Amateur radio operators have held RDF contests since the earliest days of radio, using new technologies as they evolve. If amateurs in your area hold weekend RDF contests (often called "T-hunts" or "foxhunts"), they may have RDF experience that can help you or your agency in setting up your RDF capability.

RDF is not usually a daily activity, but when the need arises, it may be urgent. Take time now to become familiar with RDF. When the unknown carrier comes on, will you be ready?

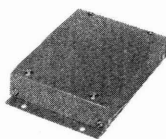


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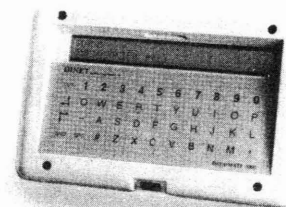
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Modulation acceptance bandwidth

By Harold Kinley, C.E.T.

The specifications on an FM communications receiver usually specify a figure for *modulation acceptance bandwidth*. What is modulation acceptance bandwidth, and what is the significance of the specification figure? We will attempt to answer that question and maybe raise some new ones.

IEEE definition

The Institute of Electrical & Electronics Engineers (IEEE) defines modulation

acceptance bandwidth as:

The selectivity characteristic of a receiver that limits the maximum permissible modulation deviation of the radio-frequency input signal that a receiver can accept, without degradation of the 12-decibel SINAD, when the radio-frequency input signal is 6 decibels greater than the reference sensitivity level. (page 4, IEEE publication standard 184-1969, *IEEE Test Procedure for Frequency-Modulated Mobile Communications Receivers*.)

Test procedure

The setup for the test procedure is

shown in Figure 1 below. This is the same test procedure setup that is used to determine the 12dB SINAD sensitivity of a receiver. To determine a receiver's modulation acceptance bandwidth, first perform

(continued on page 45)

Kinley, a certified electronics technician, is regional communications manager, South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985. He is a member of the Radio Club of America.

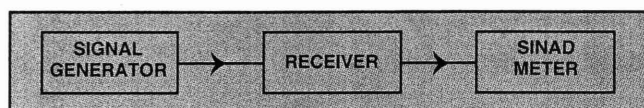


Figure 1. The setup for measuring modulation acceptance bandwidth.

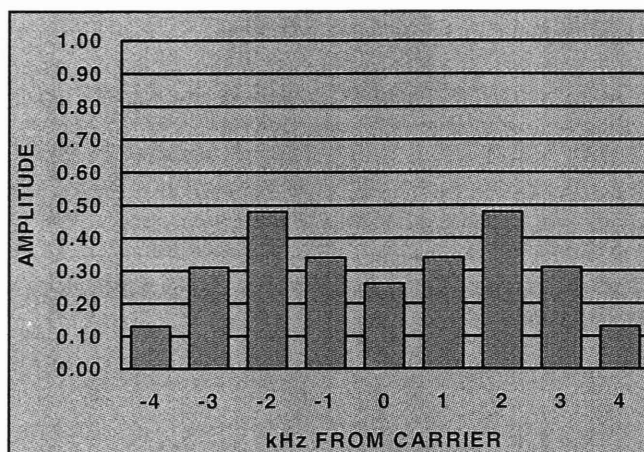


Figure 2. The significant sidebands of an FM signal modulated by a 1kHz tone at a deviation of ± 3 kHz. These sidebands account for 99.6% of the power in the sideband spectrum. Modulation index (β) = 3.

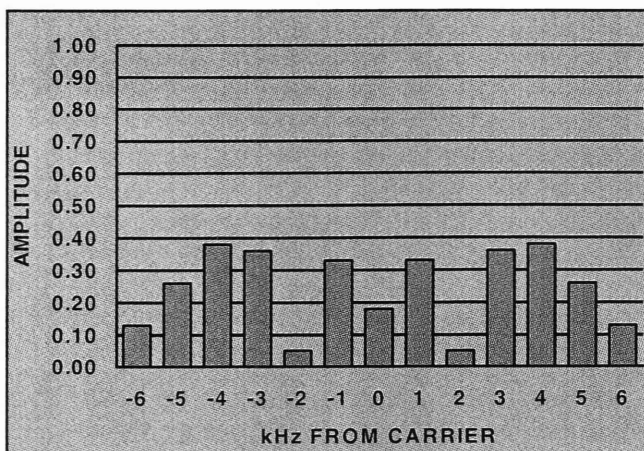


Figure 3. The significant sidebands of an FM signal modulated by a 1kHz tone at a deviation of ± 5 kHz. These sidebands account for 99.3% of the power in the sideband spectrum. Modulation index (β) = 5.

Table 1—The Bessel functions of the significant sidebands for modulation index (β) of 3, 5 and 7.

COMPONENT	AMPLITUDE		
	$\beta = 3$	$\beta = 5$	$\beta = 7$
$J_0(\beta)$	0.2601	0.1776	0.3001
$J_1(\beta)$	0.3391	0.3276	0.004863
$J_2(\beta)$	0.4861	0.04657	0.3014
$J_3(\beta)$	0.3091	0.3648	0.1676
$J_4(\beta)$	0.1320	0.3912	0.1578
$J_5(\beta)$	----	0.2611	0.3479
$J_6(\beta)$	----	0.1310	0.3392
$J_7(\beta)$	----	----	0.2336
$J_8(\beta)$	----	----	0.1280

β = MODULATION INDEX

J_0 = CARRIER

J_1 = 1ST SIDEBAND PAIR

J_n = nTH SIDEBAND PAIR

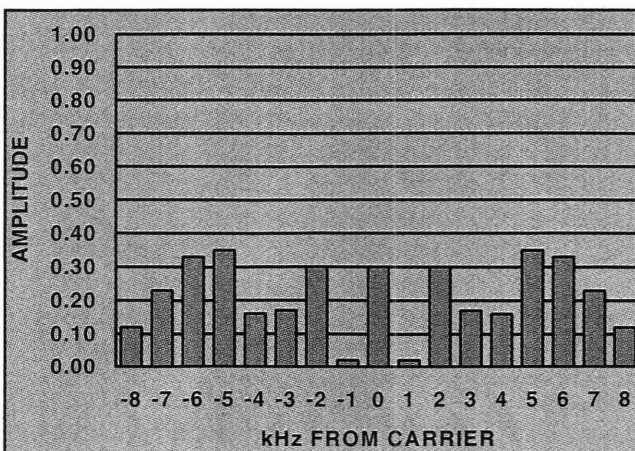


Figure 4. The significant sidebands of an FM signal modulated by a 1kHz tone at a deviation of ± 7 kHz. These sidebands account for 99.2% of the power in the sideband spectrum. Modulation index (β) = 7.

(continued from page 8)

the 12dB SINAD sensitivity test. Then, increase the signal generator level 6dB above the level required for 12dB SINAD sensitivity, or twice the microvolt level. The SINAD reading should improve greatly with the increased signal input level. Now, while observing the SINAD meter reading, slowly increase the deviation of the signal generator until the SINAD reading is degraded back to the 12dB SINAD level. At this point the deviation of the signal generator ($\pm X$ kilohertz) is the modulation acceptance bandwidth.

For narrowband FM, the *minimum* modulation acceptance bandwidth figure should be ± 5 kHz. This figure is really not enough for fully modulated signals because the modulated signal occupies a greater bandwidth. Typical receivers usually are rated at about ± 7 kHz modulation acceptance bandwidth.

With modern crystal or ceramic filters, the bandwidth usually is on the low or narrow side if the input/output impedance matching adjustments are misadjusted.

A fair test?

The test procedure described above for modulation acceptance bandwidth sometimes can be misleading and can, in some cases, lead to an unfair rating of a receiver's specification figure for modulation acceptance bandwidth. Why?

The reason is that the input signal level varies according to the *measured* 12dB SINAD sensitivity of the receiver; that is, the input signal level for determining modulation acceptance bandwidth is 6dB greater than the reference sensitivity level of the receiver.

To make a truly valid comparison of the modulation acceptance bandwidth specification between two receivers of the same model (without regard to the reference sensitivity), it would only be fair to make the measurement at the same input signal level.

You probably have seen receivers that exhibit a much better 12dB SINAD sensitivity than the rated specification. For example, a receiver with a reference sensitivity rating of $0.35\mu\text{V}$ is tested and found to have a sensitivity of $0.25\mu\text{V}$ or even $0.20\mu\text{V}$. This means that the modulation acceptance bandwidth test for that receiver is run with an input signal level of $0.5\mu\text{V}$ or $0.4\mu\text{V}$ (6dB above reference sensitivity).

Then, when the modulation acceptance bandwidth test is run at twice the measured sensitivity input level, the modulation acceptance bandwidth figure may fall short of the specification figure. In such

cases, it is better to measure the modulation acceptance bandwidth at a signal input level 6dB above the manufacturer's specification figure for sensitivity rather than 6dB above the *measured* sensitivity level.

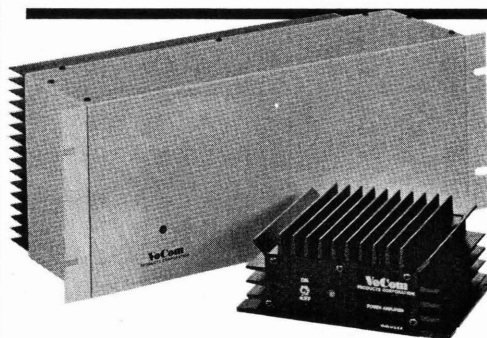
When using the modulation acceptance bandwidth test to determine the performance of a receiver or to compare the performance of two receivers, use common sense. Make the comparison on an equal

basis. Run the test in the standard way and then run the modulation acceptance bandwidth test using the same input signal level and compare the two results. This will give you a better idea of the receivers' performance under like conditions.

Modulation spectrum

Although the frequency-modulated signal has sidebands extending out to theoretical infinity, it has been determined that

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$$(2) P_{T_1} = (0.2601)^2 + 2[(0.3391)^2] + 2[(0.4861)^2] + 2[(0.3091)^2] + 2[(0.1320)^2] = 0.06765 \\ + 0.23 + 0.4726 + 0.1911 + 0.0348 = 0.99615 = 99.6\%$$

$$(3) P_{T_2} = (0.1776)^2 + 2[(0.3276)^2] + 2[(0.04657)^2] + 2[(0.3648)^2] + 2[(0.3912)^2] \\ + 2[(0.2611)^2] + 2[(0.1310)^2] = 0.0315 + 0.2146 + 0.00434 + 0.2662 \\ + 0.3060 + 0.1363 + 0.03432 = 0.99326 = 99.3\%$$

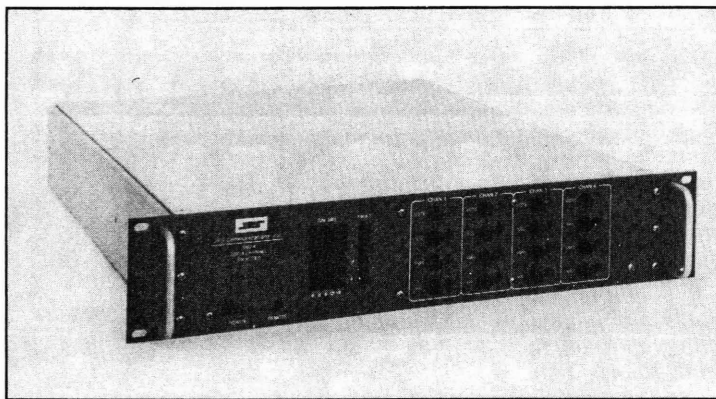
$$(4) P_{T_3} = (0.3001)^2 + 2[(0.004863)^2] + [(0.3014)^2] + 2[(0.1676)^2] + 2[(0.1578)^2] \\ + 2[(0.3479)^2] + 2[(0.3392)^2] + 2[(0.2336)^2] + 2[(0.1280)^2] = 0.0900 \\ + 0.000047 + 0.1817 + 0.0562 + 0.0498 + 0.2421 + 0.2301 + 0.1091 + 0.0328 \\ = 0.9918 = 99.2\%$$

sidebands containing 98% of the power in the modulated signal are confined to a relatively narrow bandwidth. Furthermore, if the sidebands containing at least 98% of the signal power are passed through the receiver, the signal can be reproduced with little distortion.

Figures 2, 3, and 4 on page 8 show the modulation sideband spectra of a signal frequency modulated by a 1kHz single tone at deviations of $\pm 3\text{kHz}$, $\pm 5\text{kHz}$ and $\pm 7\text{kHz}$, respectively. This represents a modulation index (β) of 3 for Figure 2; 5 for Figure 3; and 7 for Figure 4. Table 1 on page 8 shows the Bessel function for the carrier and sidebands for each of the modulation sideband spectra.

The carrier and sideband pairs in Figure 2 contain 99.6% of the total power in the frequency-modulated signal. The carrier and sideband pairs in Figure 3 contain 99.3% of the total power, and the carrier and sideband pairs in Figure 4 contain 99.2% of the total power. The percentage of power contained in the sideband pairs and carrier is calculated as shown in the box to the left. The generalized expression for the formula is shown in Equation 1 in the box. Equation 2 represents the

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calculation of total power in the sideband spectrum shown in Figure 2; Equation 3 is for Figure 3; and Equation 4 is for Figure 4.

Required bandwidth

The bandwidth required to pass the modulated sideband spectra shown in Figures 2 through 4 can be calculated from the following formula:

$$B_w = 2f_m(\beta + 1)$$

where

B_w = bandwidth in kHz

f_m = modulating frequency in kHz

β = modulation index

★ Modulation index (β) = deviation ÷ modulating tone frequency.

You can see that the formula for bandwidth holds true for the sideband spectra shown in Figures 2, 3 and 4. For example, in Figure 2 the modulation index is 3 (3kHz deviation and 1kHz tone). Substituting into the formula:

$$B_w = (2)(1)(3 + 1) = 8$$

Thus, the bandwidth is 8kHz. In Figure 2, the significant sidebands extend to ± 4 kHz, which corresponds to a bandwidth of 8kHz. This relationship can be verified in like manner for the sideband spectra shown in Figures 3 and 4.

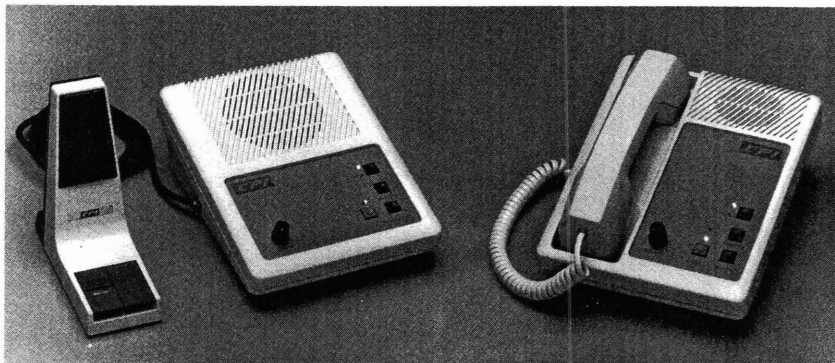
★ These illustrations show why the required bandwidth is more than just the amount of deviation of the signal. If at least 98% of the power in the modulated spectrum is passed by the receiver, the audio modulation can be reproduced with little distortion. However, when significant sidebands fall outside the bandpass of the receiver, distortion results. This phenomenon is why the modulation acceptance bandwidth specification of a receiver is so important.

That being the case, it is equally important to be able to compare the performance features of two receivers fairly in making a true performance determination. Thus, it is very useful to run the modulation acceptance bandwidth test at equal signal input levels.

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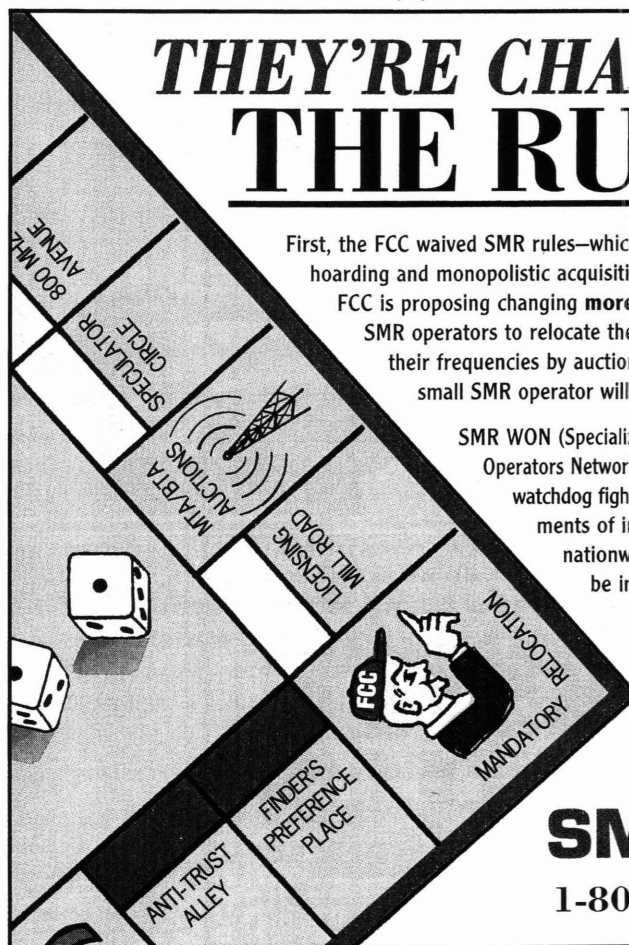
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Technically speaking

(continued from page 8)

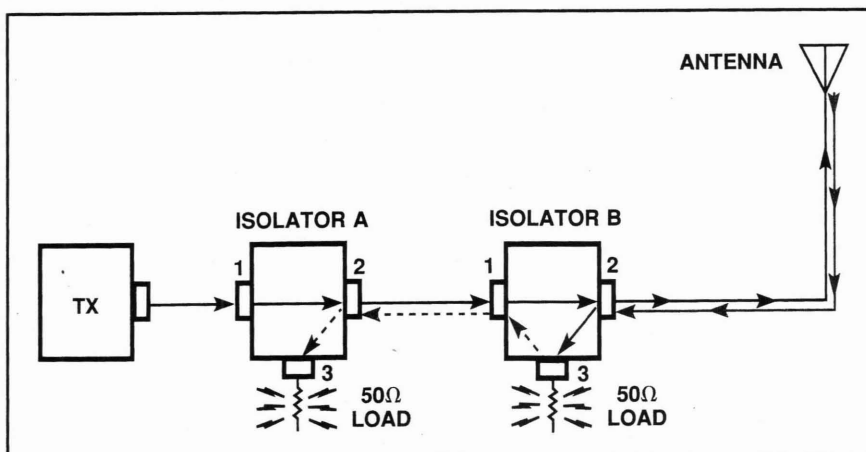


Figure 3. Two isolators can be connected in cascade to provide even greater isolation. The load resistor connected to port 3 of *isolator B* dissipates most of the reflected power, shown in red.

port 2 directly to port 3 and dissipated as heat by the dummy load at port 3. Any power reflected by the dummy load at port 3 is routed back to port 1 (broken red line) and thus back into the transmitter.

On the other hand, the RF level represented by the broken red line going to port 1 normally is much smaller than the RF level dissipated by the dummy load at port 3. Thus, most of the reflected power is dissipated as heat in

the dummy load and not returned to the transmitter through port 1.

Single-section vs. dual-section

When more isolation is desired than a single-section isolator can provide, two or more isolators can be connected in cascade to achieve the desired isolation.

Dual-section and even triple-section isolators are available from most isolator manufacturers. There are advantages and disadvantages to cascading single-section isolators or using dual-section or triple-section isolators or a combination.

Figure 3 to the left shows how two separate single-section isolators may be connected in cascade to provide additional isolation. If both isolators are identical, the isolation provided by two connected in cascade is twice that of one alone. At the same time, the insertion loss is twice (or more) that of a single section.

If there is a severe impedance mismatch at port 2 of isolator B, most of

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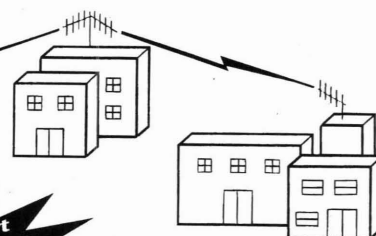
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Technically speaking

the reflected power is directed into the load at port 3 of isolator B and is dissipated as heat. Because the load at port 3 of isolator B should provide a good impedance match, little power is reflected back into port 3 of isolator B and on to isolator A.

Nevertheless, any reflected power entering isolator A from isolator B is directed to the load at port 3 of isolator

A. The power dissipation of the load at port 3 of isolator A is small because the brunt of the reflected power is taken by the load at isolator B. Thus, the load connected to isolator A need not be as large as the load connected to isolator B.

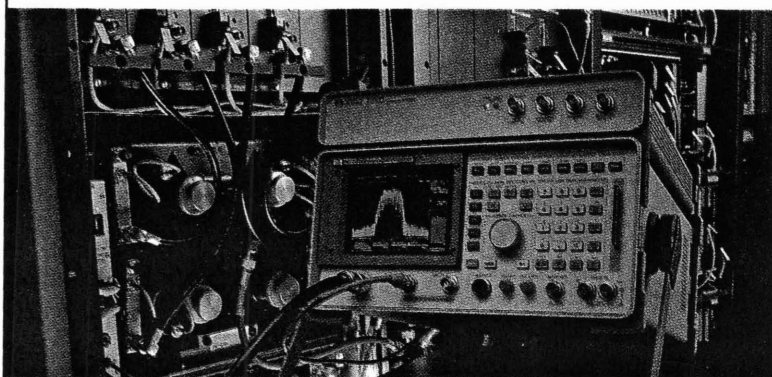
A dual-section isolator can provide the same isolation as two single-section isolators in cascade and cause

less insertion loss. In addition, a dual-section isolator costs less than two single-section isolators. Still, two single-section isolators are not *twice* as expensive as an equivalent dual-section isolator.

Another advantage of the dual-section isolator is that it is less cumbersome to connect because the interconnecting cable between the output of one isolator and the input of the next is eliminated, moreover, the dual-section isolator requires less space.

One advantage of using two single-section isolators instead of one dual-section isolator is the repair-and-

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A dual-section isolator can provide the same isolation as two single-section isolators in cascade and cause less insertion loss. In addition, a dual-section isolator costs less than two single-section isolators.

replacement factor. If one section of a dual-section isolator goes bad, the entire isolator must be replaced. Two single-section isolators can be tested and replaced separately, reducing replacement cost.

A further advantage of using two single-section isolators is to provide better dissipation because the heat generated is dissipated over a greater surface area.

Simple mathematical analysis

Refer to Figure 3.

Suppose that the two isolators, A and B, are identical single-section isolators with an isolation of approximately 20dB and an insertion loss of 0.35dB. Assume that the power input from the transmitter to port 1 of isolator A is 100W.

Technically speaking

Further assume that the return loss at the output of isolator B (port 2) is 10dB. According to Table 1 to the right this return loss is equivalent to a VSWR of 1.93:1. The forward power at the output of isolator B (port 2) is 0.7dB below 100W. (This 0.7dB is the combined insertion loss of the two isolators.) This insertion loss reduces the forward power to 85.1W.

Because the return loss is 10dB, the reflected power level is 8.5W at port 2 of isolator B. This reflected power is directed to port 3 of isolator B, where most of it is dissipated in the load.

This is a small loss—approximately 0.35dB—from ports 1 to 3 of isolator B, the same as the insertion loss from ports 1 to 2. This loss reduces the power appearing at port 3 to 0.35dB



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Table 1—Return loss vs. VSWR

Return Loss	VSWR
0	∞
1	17.39
2	8.72
3	5.85
4	4.42
5	3.57
6	3.01
7	2.62
8	2.32
9	2.10
10	1.93
11	1.78
12	1.67
13	1.58
14	1.50
15	1.43
16	1.38
17	1.33
18	1.29
19	1.25
20	1.22

below 8.5W or 7.8W. Assuming the load at port 3 of isolator B presents a good match with a return loss of 20dB (VSWR of 1.22:1), the power reflected back into port 3 is 0.07W or 78mW and is directed on to port 1 of isolator B.

The reflected power appearing at port 1 of isolator B is 0.35dB below 78mW or 72mW and passes on through the interconnecting cable to port 1 of isolator A to port 3 of isolator A. The reflected power appearing at port 3 of isolator A is 0.35dB below 72mW or 66.4mW.

Again, assuming the load at port 3 of isolator A presents a good match with a return loss of 20dB (VSWR of 1.22:1), the power reflected back into port 3 of isolator A is 0.664mW or 664μW.

Thus, at isolator A, a forward power of 100W enters port 1, and a reflected power of 612.6μW leaves port 1. The return loss is $10\log(100/0.000612) = 52.1\text{dB}$ or a VSWR of 1.005:1. This example shows how the isolator maintains a constant impedance match to the transmitter.

Technically speaking

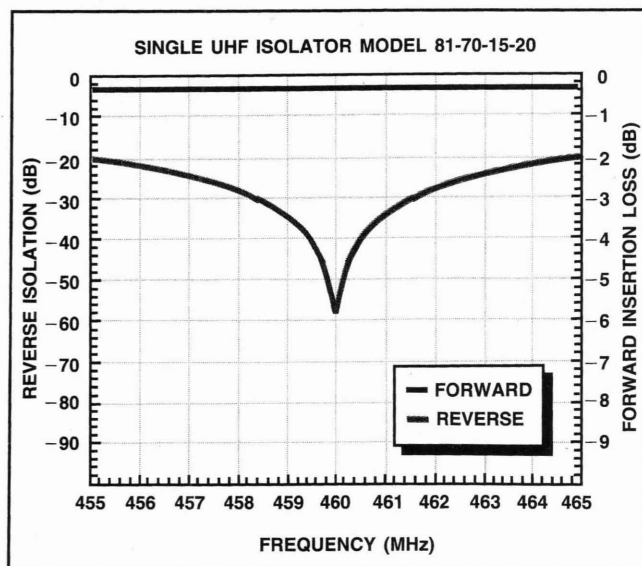


Figure 4. This graph shows both the insertion loss and the isolation with respect to frequency for a single-section isolator. (Courtesy TX RX Systems.)

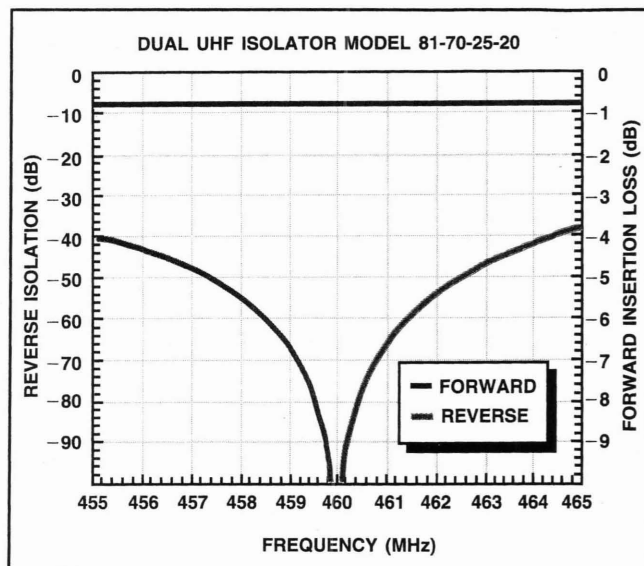


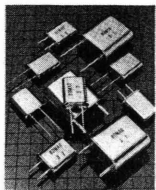
Figure 5. This graph shows the insertion loss and the isolation with respect to frequency for a dual-section isolator. (Courtesy TX RX Systems.)

Notice that the return loss at the transmitter output is 52.1dB while the return loss at the output of port 2 of

isolator B is 10dB. The return loss at the transmitter should equal the combined isolation figures of the isolators

plus the return loss at the output port of the final isolator plus the combined insertion loss figures of the isolators.

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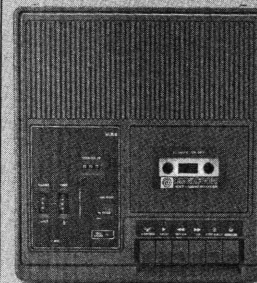
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Technically speaking

In Figure 3, this is $20.7\text{dB} + 20.7\text{dB} + 10\text{dB} + 0.35\text{dB} + 0.35\text{dB} = 52.1\text{dB}$.

The isolation provided by the isolators in the example can be calculated using the formula:

$$\text{Isolation (dB)} = 10 \log(P2/P1)$$

where

P2 = reflected power at port 2

P1 = reflected power at port 1

In the example, the reflected power at port 2 of isolator B is 8.5W (P2) and the reflected power at port 1 is 72mW. Substituting into the formula:

$$\text{Isolation dB}$$

$$= 10 \log(8.5/0.072)$$

$$= 10 \log(118.055)$$

$$= 20.7\text{dB}$$

It is important that the load be properly matched to the isolator port. Any power reflected back into port 3 places an additional power dissipation requirement on the isolator.

The same formula can be applied to isolator A in the example, and the isolation figure comes out the same: 20.7dB. Typical isolation figures for single-section units are better than 30dB, so the performance would exceed that of the example.

Figure 4 on page 48 graphs isolation vs. frequency of a typical single-section isolator. The insertion loss vs. frequency also is shown.

Figure 5 on page 48 is a similar graph for a dual-section isolator. No-

tice that the isolation provided by the dual-section isolator is much greater than that provided by the single-section isolator. Notice, too, that the insertion loss of the dual-section isolator is greater.

Heat: The enemy

Heat is the number one enemy of isolators.

As long as isolator temperature is kept within the specified operating range, typically -30°C to $+60^{\circ}\text{C}$ or -22°F to $+140^{\circ}\text{F}$, the isolator should perform as specified. When the isolator is operated outside this temperature range, disaster can result. Here are some recommendations for avoiding such a disaster:

(1) Avoid exceeding the specified input power rating.

(2) Use a good-quality dummy load on the termination port.

It is important that the load be properly matched to the isolator port. Any power reflected back into port 3 places an additional power dissipation requirement on the isolator.

Further, the load should have good, thermally efficient radiating fins to radiate the heat instead of conducting it back into the isolator.

(3) Do everything possible to minimize the power reflected back into port 2 from the antenna, feedline and duplexer. Keeping the reflected power low sends less power into the dummy load and reduces the heat generated there.

(4) If it is necessary for the dummy load to dissipate much power, then consider physically separating the load from the isolator by using a well-shielded connecting cable. This separation reduces the heat transferred by conduction from the load to the isolator.

(5) Avoid placing the isolator in a high-ambient-temperature location.

If necessary, use fans to force air over the isolator to keep the temperature within safe operating limits.

(6) Use common sense.

Next month, we take a closer look at various isolator specifications and how to operate the devices within safe limits.



Isolators—the practical view, part 2

By Harold Kinley, CET

In last month's column, some basic concepts of isolators were presented from a practical point of view.

This column includes some additional information about isolator use.

Input power vs. return loss

One isolator specification figure is input power rating.

If the reflected power is negligible (high return loss), then the isolator can be operated safely at full input power rating without exceeding its internal dissipation.

On the other hand, when the reflected power becomes significant, it increases the heating within the isolator and exceeds its maximum rated dissipation. To maintain a safety factor, the input power should be reduced to offset the increased dissipation caused by the reflected power.

Refer to Figure 1 above. Assume that both of the isolators shown in the figure are rated for 100W input power. The insertion loss is 0.35dB. The power input to isolator A is 100W (full

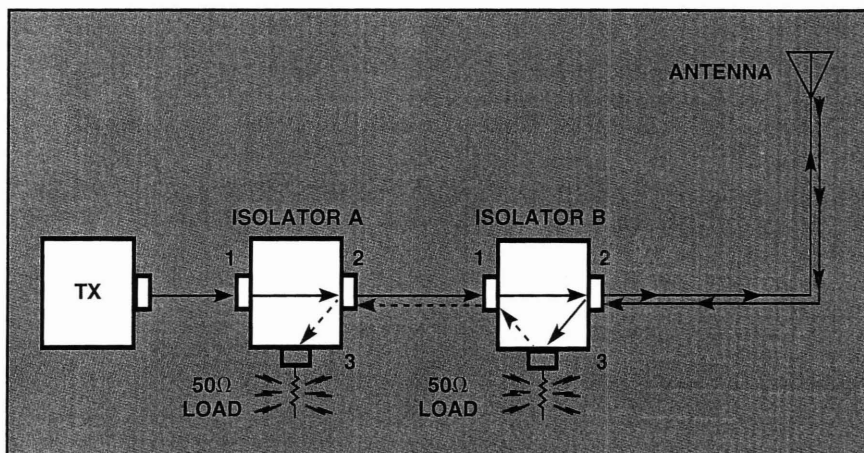


Figure 1. Two isolators can be connected in cascade to provide even greater isolation. The load resistor connected to port 3 of isolator B dissipates most of the power, shown in red.

rated power).

Because the insertion loss is 0.35dB, the output power at port 2 is 92.3W. Thus the power dissipated within the isolator is $100W - 92.3W = 7.7W$. To play it safe, 7.7W should be considered to be the isolator's maximum safe dissipation. As long as no reflected power enters the isolator and the input power does not exceed 100W, the in-

ternal dissipation will not exceed 7.7W.

When the load connected to the isolator output port is badly mismatched (low return loss, high VSWR), significant power is returned to port 2 and on to port 3 where it is "dumped" into the load. If the return loss is 3dB, for example ($VSWR = 5.85:1$), 50% of the output power from port 2 of the isolator reflects back into port 2 and travels to port 3 into the load.

In passing from port 2 to port 3, this reflected power causes increased dissipation within the isolator. If the isolator is operating at the 100W maximum input power, the output power from port 2 is 92.3W. The reflected power is 46.2W. The reflected power appearing at port 3 is 42.6W. The extra dissipation caused by the reflected power is $46.2W - 42.6W = 3.6W$.

Because of the reflected power, the isolator must dissipate $7.7W + 3.6W = 11.3W$. To bring the internal dissipation back to a safe level, the input power should be reduced to compensate for the reflected power.

(continued on page 50)

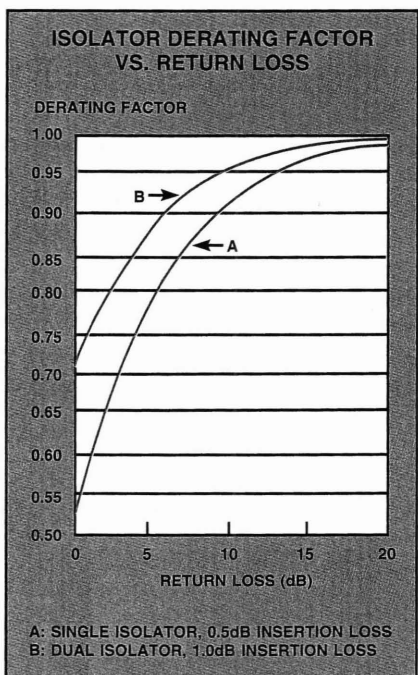


Figure 2. Use this graph to determine the derating factor to be applied to the input power for a given amount of return loss.

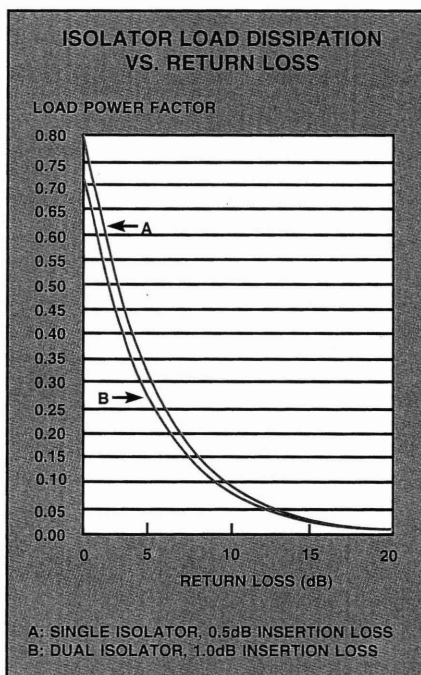


Figure 3. Use this graph to determine the required isolator load dissipation for a given amount of return loss.

Kinley is a certified electronics technician with the South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985.

Technically speaking

(continued from page 8)

The formula for finding the proper derating factor is:

$$F_d / (F_d + R_d)$$

where

F_d = forward power dissipation (watts)

R_d = reflected power dissipation (watts)

Because finding F_d and R_d is a bit tedious in itself, the following formula offers a quicker way to find the solution.

For single-section isolators:

$$D_f = \frac{1}{1 + \text{antilog} \left[\frac{(L+R)}{10} \right]}$$

where

D_f = derating factor

L = insertion loss

R = return loss

Be certain to enter L and R as negative values because they represent a loss.

$$D_f = \frac{1 + \text{antilog} \left[\frac{L}{20} \right]}{1 + \text{antilog} \left[\frac{L}{20} \right] + \text{antilog} \left[\frac{L+R}{10} \right]}$$

For dual-section isolators:

Again, be sure to enter L and R as negative values.

Figure 2 on page 8 can be used to find the derating factor for single-section isolators with 0.5dB insertion loss and for dual-section isolators with 1.0dB insertion loss.

For example, a certain single-section isolator has an input power rating of 150W under well-matched operating conditions (high return loss, low VSWR). Suppose that the return loss drops to 3dB. According to curve A on the graph in Figure 2, the derating fac-

tor for a return loss of 3dB is 0.69.

Thus, the normal input power rating should be reduced by a factor of 0.69. The reduced input power level should be no more than 69% of the specified maximum or $0.69 \times 150W = 103.5W$. This reduction compensates for the increased heating effect of the reflected power. For dual-section isolators, use curve B on the graph.

Dummy load size

Technically speaking, the dummy load connected to port 3 must be able to dissipate the full transmitter power if a worst-case mismatch occurs at the output (port 2). If such a worst-case mismatch occurs, almost all of the transmitter power will be "dumped" into the dummy load. If the load is incapable of dissipating this much power, the dummy load and possibly the isolator will be damaged from overheating.

Such worst-case mismatches indeed can occur, but usually what the designer wants to protect against is an-

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Technically speaking

tenna damage that causes the VSWR to soar and return loss to drop.

For example, if the isolator output is connected to an antenna through a feedline with a 1.5dB loss and a worst-case mismatch occurs at the antenna, the return loss at the feedline input will be 3.0dB. Thus, half of the output power will be dumped into the dummy load and dissipated as heat.

Usually, the size of the load is chosen to be approximately 50% of the transmitter output power; nonetheless, there are no hard and fast rules. Obviously, you would not elect to use a 200W dummy load if the transmitter were rated at 50W! Be conservative, but do not take it to the extreme. Use the common-sense approach.

The graphs in Figure 3 on page 8

can be used to determine the load requirements for a given amount of return loss. Curves are given for both single- and dual-section isolators.

As an example, suppose the return loss is 5dB. For a single-section isolator, the load requirement would be approximately 25% of the transmitter power. The following formula can be used to calculate the load requirement for a single-section isolator:

$$L_f = \text{antilog} \left[\frac{R+2L}{10} \right]$$

where

L_f = load factor

R = return loss in dB

L = insertion loss in dB

For dual-section isolators, the formula is:

$$L_f = \text{antilog} \left[\frac{R+1.5L}{10} \right]$$

Be sure to enter R and L as negative values.

Harmonics

In *Technical Papers* published by EMR, Phoenix, Bill Lieske writes:

"When RF power is conducted by the ferrite junction, harmonic responses are likely in most designs, and it is found that, possibly due to impurities or non-linearities in the ferrite mix and other phenomena, harmonic energy produced by a transmitter may be conducted through an isolator and, in some cases, weak harmonics can be generated in the ferrite material itself."

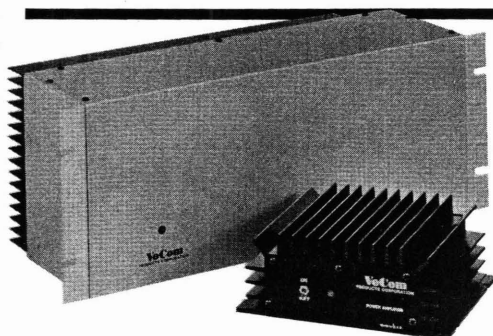
For that reason, it is important to use a harmonic filter between the isolator output and the antenna. (See Figure 4 on page 54.)

Moreover, if a *pass* cavity is connected between the isolator output and the antenna (as shown in Figure 5 on page 54), the harmonic filter is unnecessary. Figure 5 represents a typical repeater setup.

Isolator testing

After installing an isolator, you should confirm that it provides the iso-

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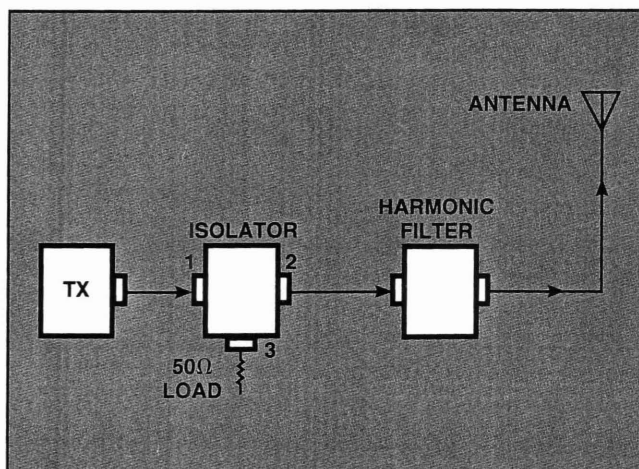


Figure 4. To suppress any harmonics caused by the isolator, a harmonic filter is used between the isolator output port and the antenna.

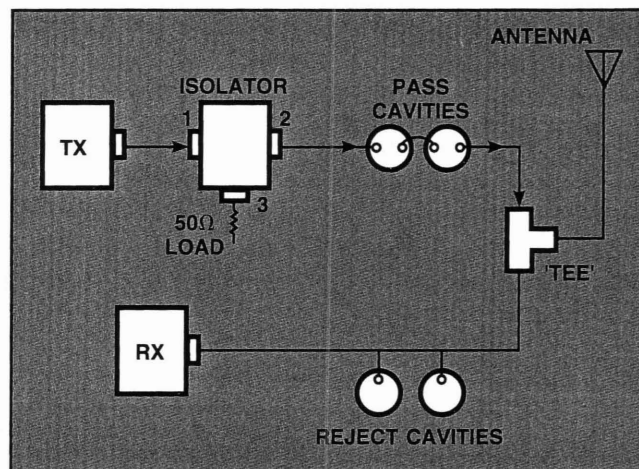


Figure 5. When a bandpass cavity follows the isolator, no special harmonic filter is needed. This configuration is typical of a repeater setup.

lation that the manufacturer specifies.

A simple way to measure isolation is shown in Figure 6 to the right. Set the signal generator to 0dBm, and measure the signal level (at the input port) on the spectrum analyzer.

First, feed the signal generator output directly into the spectrum analyzer to establish the 0dBm reference on the spectrum analyzer display. If the manufacturer specifies the isolation figure

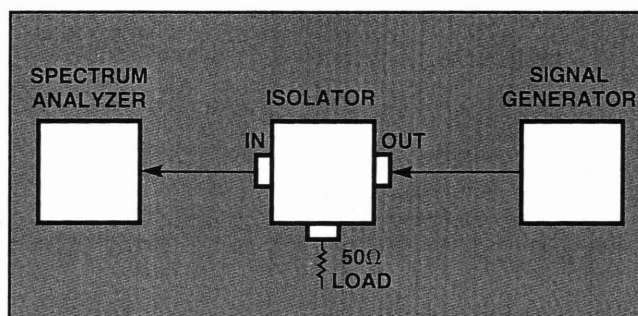


Figure 6. This setup is used to check the isolator's isolation figure. Notice that the signal is fed into the isolator output port while the measurement on the spectrum analyzer is taken at the input port. The difference between the signal generator level and the level measured on the spectrum analyzer is equal to the isolator's isolation figure.

to be 35dB, then the signal level measured on the spectrum analyzer should be at or below -35dBm. If you measure an isolation figure radically different from the specified figure, something is wrong, either in the isolator or in your test equipment.

The insertion loss should be checked, too.

Figure 7 on page 56 shows a simple way to measure the insertion loss. At (a), the forward power entering the input port is measured and re-

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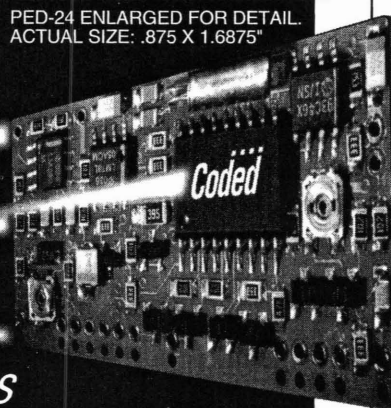
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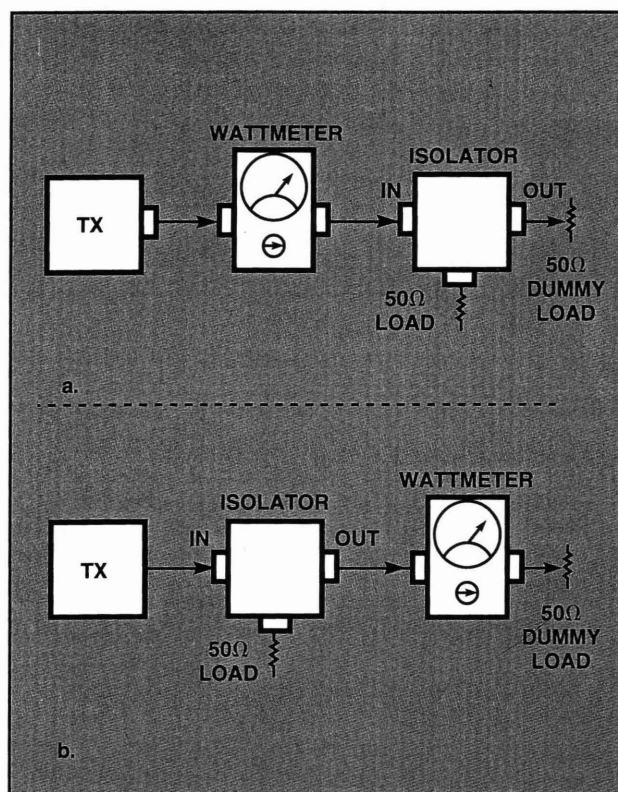
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Technically speaking

Figure 7. This setup can be used to measure the isolator's insertion loss. At (a) the wattmeter measures the input power to the isolator. At (b) the wattmeter measures the isolator output power. The insertion loss then is determined by the formula $10 \log(P_i/P_o)$.



recorded as P_i . At (b), the forward power exiting the output port is measured and recorded as P_o . The insertion loss then is found by using the formula:

$$\text{Insertion loss} = 10 \log(P_i/P_o)$$

For example, if the input power P_i is 100W and the output power P_o is 90W, the insertion loss is $10 \log(100/90) = 10 \log(1.111) = 0.46\text{dB}$. Any radical departure from the specified figure should be investigated more closely.

Simplex base station hookup

When hooking up an isolator to a simplex base station, be certain to insert the isolator in the line between the transmitter output and the antenna relay.

If the isolator is inserted beyond the antenna relay (between the antenna relay and the antenna), the antenna will be effectively isolated from the receiver input, rendering the receiver virtually inoperative. (See Figure 8 on page 58.)

Proposed standard

Lieske offers the following information as a recommended standard to be used by transmitter site managers in controlling intermodulation interference among users:

"An RF isolator shall be provided immediately following each transmitter and shall provide a minimum of 30dB of isolation. In addition, the isolator shall be followed by a cavity resonator, harmonic, bandpass or lowpass filter to ensure that all conducted or generated harmonics are at least 100dB below the transmitter carrier power measured at the antenna system feedpoint.

"In the event that an intermodulation product resulting in interference to any on-site receiver is traced to a given transmitter, additional isolators or isolator sections shall be provided as required to suppress intermodulation products to a non-interfering level.

"All isolators shall maintain the minimum of isolations herein defined throughout the ambient temperature ranges to be found at the site and up to the maximum duty cycle of the radio transmitter concerned.

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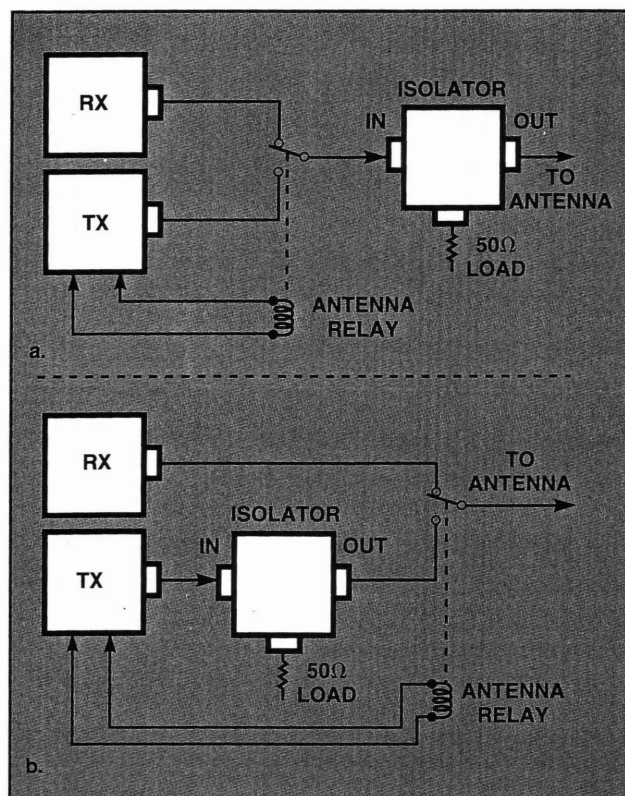


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Technically speaking

Figure 8. At (a), the isolator is incorrectly inserted in the common antenna line to the transmitter and receiver. If this setup were used, little signal could reach the receiver input because the antenna is, in effect, isolated from the receiver. At (b), the isolator is installed between the transmitter output and the antenna relay. In this arrangement, the isolator is not between the antenna and the receiver input; thus, the received signal is not isolated from the receiver input.



interference still be found after the application of up to three cascaded isolators or three isolator sections, further reduction of intermodulation products shall be accomplished through (a) the addition of cavity filters, (b) through separation of antennas to reduce couplings, (c) a combination of (a) and (b), or (d) the adjustment of transmitter power output powers as needed to reduce the interference to an acceptable level."

It is hoped that this column, along with last month's, has answered some of your questions about isolators.

For the "purists," a computer program that computes the derating factor and the load requirement of typical isolators (single and dual) is available from the author at P.O. Box 15178, Spartanburg, SC 29302. The cost is \$5, including shipping and handling. The program is available only on a 5.25-inch floppy diskette in IBM format.

Additional information about isolators is available from various manufacturers.

Two manufacturers that can furnish additional tutorial information about isolators are: EMR, 22402 N. 19th Ave., Phoenix, AZ 85027; and Tx Rx Systems, 8625 Industrial Parkway, Angola, NY 14006.

References

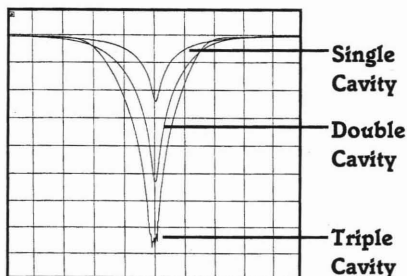
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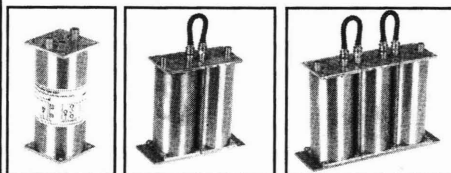
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How mobile radio fading affects data transmission

Part 1—An intuitive understanding of RF fading is possible without complex mathematics and statistics. Radio signal reflections can have positive effects as easily as negative effects. Packet length makes a difference.

By Barry L. Dorr

Most of us in the mobile communications industry are aware of the effects of RF signal fading. As mobile data systems become more commonplace, these effects are becoming more important. Multipath, delay spread, coherence bandwidth, Rayleigh fading and Doppler frequency are terms used to describe signal fading. The following information explains the basics of fading and the terminology. A second article describes how well-designed data modems and systems allow data transmission over fading channels.

The basics

Radio waves are reflected by stationary objects (e.g., buildings) as well as moving objects as shown in Figure 1 above right.

Dorr is an independent consultant in Carlsbad, CA. He specializes in modem design, communication system design and digital signal processing algorithm development. Tel. 619-944-3968.

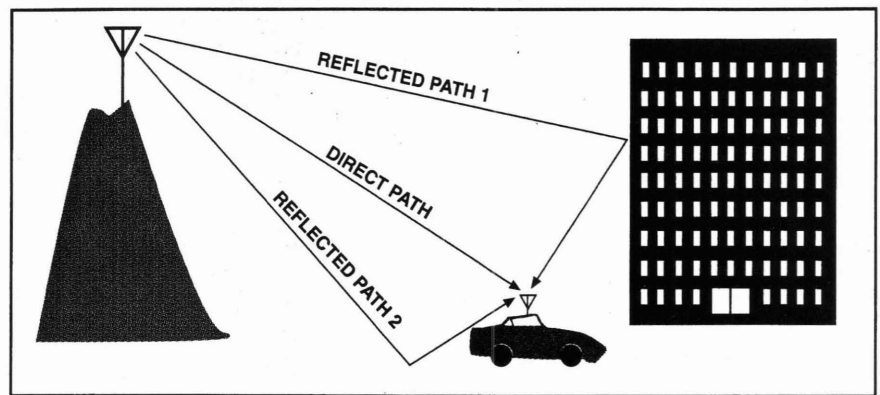


Figure 1. A specific example of multiple reflected and attenuated signals arriving at a receiver. A direct ray and two reflections are shown. A direct ray and six reflections are used for the example in the text.

Reflections have positive and negative effects: On the positive side, a reflection may allow reception when the direct, or line-of-sight, path is blocked. On the negative side, receivers often pick up the line-of-sight signal and reflected versions of the signal. Compared to the line-of-sight sig-

nal, reflected signals usually are delayed slightly and attenuated. Because multiple reflected signals enter the receiver, the term *multipath fading* often is used to describe this phenomenon.

Different frequencies and propagation conditions produce various multipath

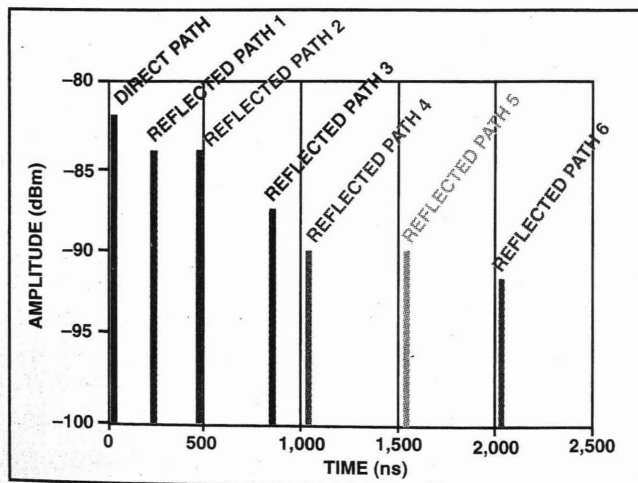


Figure 2. Received signal envelope amplitude vs. arrival time for the reflected signals shown in Figure 1. The direct-path signal reaches the receiver at time = 0.

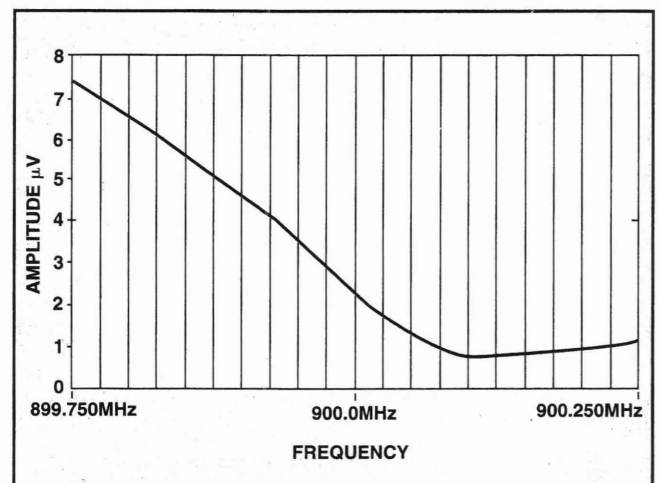


Figure 3. Received signal amplitude vs. frequency for reflected signals shown in Figure 1. Note that the frequency increments are spaced by 25kHz.

effects. For example, a busy street with surrounding buildings would tend to have many reflections, whereas an unobstructed radio path in the desert would have few. A term used to quantify multipath effects is *delay spread*. Delay spread is the amount of time required for most of the reflections of a signal to reach a receiver.

To illustrate further, consider a specific example of a static fading condition. This example shows a *specific* case of a *random*

process. (The concept of fading is well-illustrated by this example; nevertheless, communication system designers normally use complex statistical analysis to describe these effects.) In Figure 1, a short burst of RF energy is transmitted from the base station. The envelope of the signal seen by the receiver is shown in Figure 2 on page 30. Note that all of the energy is received within a 2,500ns (2.5 μ s) window. To determine delay spread accurately, many data

points (obtained with the vehicle in different places) must be averaged. A typical value of delay spread for a 900MHz signal in an urban area is 2.5 μ s.

Multipath fading also limits the bandwidth of data signals that can be transmitted over a channel. When the delayed and attenuated reproductions of the line-of-sight signal are received, some add constructively and some cancel each other by adding destructively. Suppose that the frequency of the unmodulated carrier from the base station in Figure 1 was varied and the received signal envelope was measured at the mobile. The result would be the plot shown in Figure 3 on page 30. Data signals are composed of many frequencies. Figure

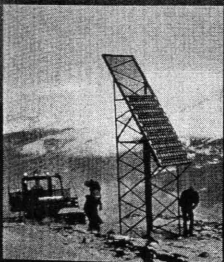
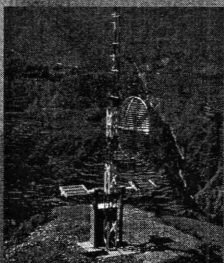
For a given channel, the bandwidth over which the magnitude and phase stay relatively constant is the coherence bandwidth.

3 shows that any data signal wider than several hundred kilohertz probably would be distorted by this channel. For a given channel, the bandwidth over which the magnitude and phase stay relatively constant is the *coherence bandwidth*. As with delay spread, to determine coherence bandwidth accurately, many data points (obtained with the vehicle in different places) must be averaged. A typical value for coherence bandwidth for an 900MHz signal in an urban area is 150kHz. A wider data signal probably would need some *equalization* to flatten the channel.

Marks on Figure 3's horizontal axis represent 25kHz increments. The figure shows that coherence bandwidth normally is not a problem with channels as they currently are assigned at 900MHz. For higher bit-rate data services, (such as personal communications services [PCS]) coherence bandwidth will become one of the primary bit-rate constraints because equalizers will raise the cost of subscriber equipment significantly.

If the vehicle in Figure 1 were moved to another location, the curves in Figures 2 and 3 would change somewhat. If the vehicle was moved to many locations and the received signal strength was measured at each location, the data would have a Rayleigh distribution as

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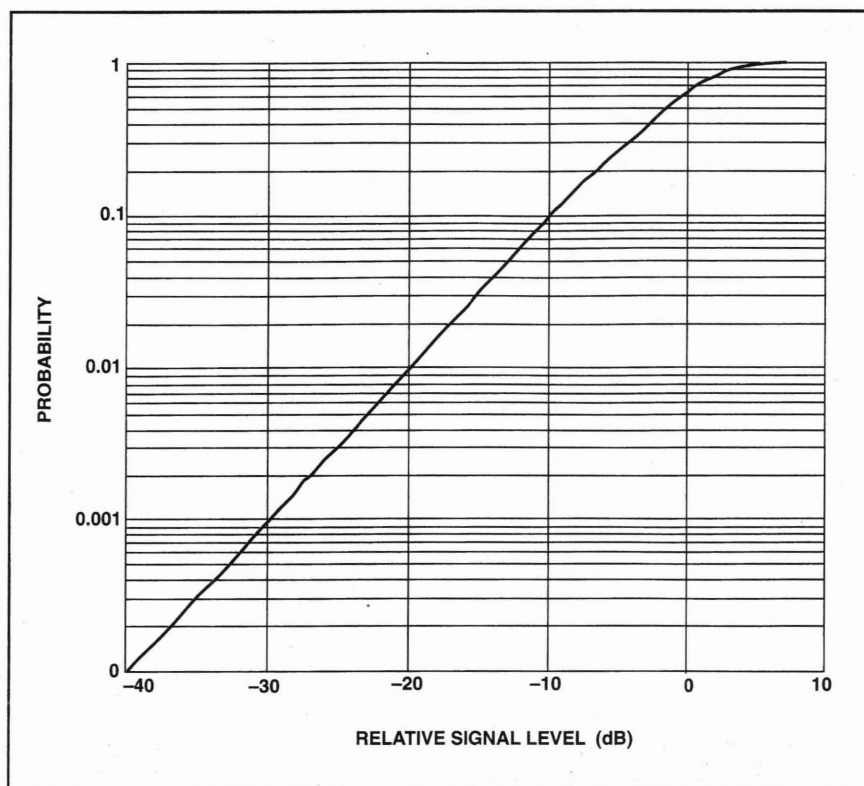
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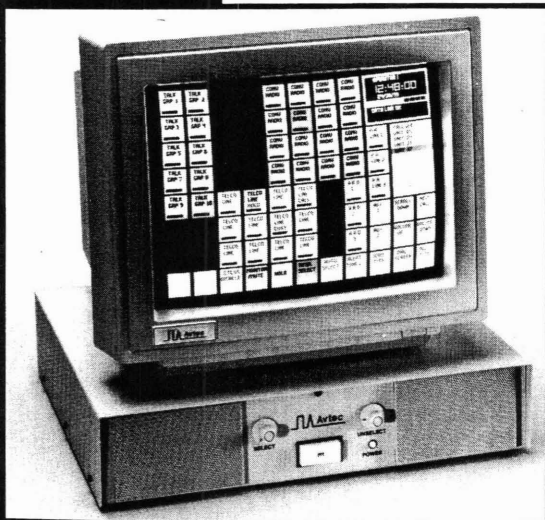
shown in Figure 4 to the left.

Moving-vehicle fading effects

As the vehicle in Figure 1 moves past reflective objects, reflections appear and disappear. Also, the lengths of the reflected paths change. Another effect is that Doppler shift causes the different paths to arrive at slightly different frequencies. The overall result is that the amplitude and phase of the received signal are constantly changing. If the base was sending an unmodulated carrier, the received signal level would vary randomly. If many samples of the received signal envelope were measured, they would follow the Rayleigh distribution shown in Figure 4. The received signal level fluctuations occur mostly at the Doppler frequency, which is a function of the RF frequency and the vehicle speed as shown in the following equation.

Figure 4. Cumulative Rayleigh distribution. From this curve you can determine the probability that the instantaneous received signal level (RSL) will be attenuated below the mean RSL by a given amount. As an example, the probability of the received signal being less than 20dB below the average level is 1%.

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$$f_d = \frac{v}{\lambda}$$

where:

f_d = Doppler frequency

v = vehicle speed in meters-per-second

λ = carrier wavelength in meters

Putting it all together

As an example, for a 900MHz carrier and a 40mph vehicle speed, the Doppler frequency is 53.64Hz. The transmitted signal is an unmodulated carrier. A sample plot of the receive signal level at the mobile is shown in Figure 5 to the right. As expected, the fluctuations roughly correspond to the Doppler frequency. In addition, note that the fluctuations all have different depths.

Figure 5 provides insight about data transmission under these circumstances. As a rough approximation, suppose that if the received signal level was below -110dBm, the data modem could not decode bits; above this threshold, no errors

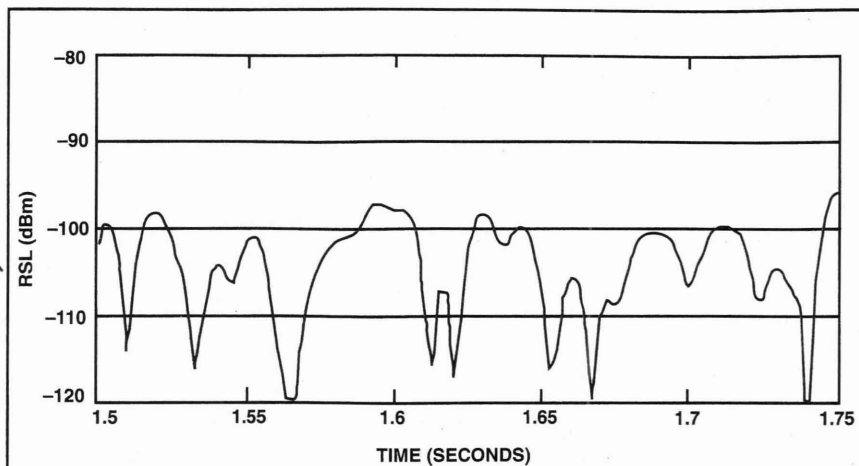


Figure 5. Plot of received signal level vs. time for a vehicle traveling 40mph. The carrier frequency is 900MHz, and the mean received level is -101dBm.

would occur. If no error correction were used, then packets exceeding 50ms probably never would be decoded by the receiver. Conversely, extremely short packets would arrive reliably. This difference explains why some vehicular data modems use interleavers and powerful coding techniques and other modems achieve high

performance with no error correction whatsoever.

Next: Several practical methods used by data modems for passing data in the presence of fading, and practical methods for establishing system performance.



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Pre-distortion simplifies mobile data integration

Interfaces between data modems and existing voice-grade radios are more efficient with pre-distortion signal processing. Thus, a larger number of radio models can be used for 4,800bps and 9,600bps operations.

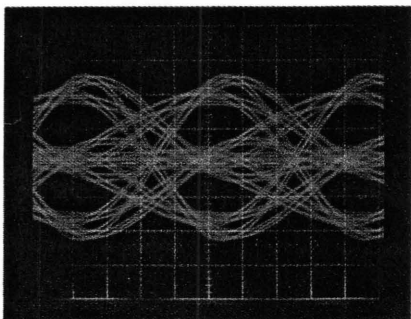


Photo 1. An eye pattern distorted by the Mitrek circuitry.

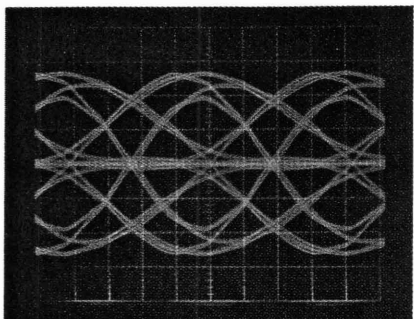


Photo 2. An eye pattern of a pre-distorted signal.

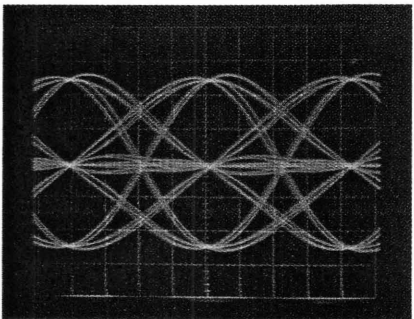


Photo 3. An eye pattern of a pre-distorted signal after it passes through the Mitrek circuitry. Note how the resulting signal is symmetrical and how it converges to three distinct points.

By Barry Dorr

Pre-distortion is a signal processing technique that yields more efficient interfaces between data modems and existing voice-grade radios.

A result of this technology is that a greater number of radio models can be used for 4,800bps and 9,600bps operation, and fewer radio modifications are required. Thus, users can send high-speed data while retaining their investment in mobile radio and base station equipment.

How pre-distortion works

As a general rule, higher data rates occupy more audio bandwidth than lower data rates, requiring radios with a wider audio bandwidth. Audio group delay must remain relatively flat over this bandwidth. At 4,800bps data rates, voice-grade radios normally do not meet the requirements for bandwidth and group delay.

System integrators traditionally solve this problem by modifying the radio so that the transmitted signal is sent directly to the voltage-controlled oscillator (VCO) input. When this method is not possible because of low impedance, low distortion or the lack of a good audio summing point, the radio model is rejected for use at high baud rates.

The solution is to select a convenient transmit interface point and digitally

Dorr is a senior engineer and group leader at Coded Communications, Carlsbad, CA. He has B.S.E.E. and M.S.E.E. degrees with an emphasis in communications and signal processing. The technology discussed in this article is incorporated in Coded's IQmodem product.

pre-distort the signal in the modem.

As an example, the Motorola Mitrek radio's VCO input has a 100Ω input impedance, which is too low for a low-power data modem. The radio has an interface point that has high impedance and that allows summing of the voice signal, but it does not provide a satisfactory interface at 4,800bps.

The audio magnitude and group delay of the transmit circuitry between the selected interface point and the VCO input are shown in Figure 1 on page 37. As with most radios, the Mitrek's high-

As a general rule, higher data rates occupy more audio bandwidth than lower data rates, requiring radios with a wider audio bandwidth.

frequency audio response is limited to reduce noise. Although it is not shown by the graph, the low end of the radio's audio frequency response is limited by capacitive coupling in the audio path.

The audio spectrum of the 4,800bps Nyquist-filtered bipolar (NFB) signal is shown in Figure 2 on page 37. Because the signal has no low-frequency component, the low-frequency limitation due to capacitive coupling is not a problem.

The high-frequency limitation, how-

Definitions:

Digital signal processing (DSP) — The processing of analog signal using digital techniques.

Eye pattern — The waveform pattern of a modulated digital signal as seen on an oscilloscope. The vertical display is driven by the information-carrying waveform, synchronized by the system clock.

Group delay — A type of distortion caused by different frequencies that take different amounts of time to pass through a filter.

Nyquist-filtered bipolar (NFB) — A signaling technique commonly used at 4,800bps. The most salient characteristic of this scheme is that it has no dc component. Therefore, an NFB signal passes through capacitively coupled channels.

Transfer function — A function that describes the magnitude and group delay response for a linear filter.

Voltage-controlled oscillator (VCO) — The fundamental element of an FM transmitter. The VCO converts audio-level signals into deviations of an RF carrier's frequency.

ever, causes distortion. Photo 1 on page 34 is an eye pattern distorted by the Mitrek circuitry. The distortion it represents is unacceptable.

The signal will be passed by the Mitrek if it is pre-distorted so that the transmit circuitry *actually is required*

...the transmit circuitry actually is required for a clean signal.

for a clean signal. This is done as follows:

If the transfer function of the transmit circuitry is $H_d(s)$ and the desired transfer function of the modem and processing circuitry is $H_n(s)$, then the digital filter required for the Mitrek can be expressed as $H_n(s)/H_d(s)$.



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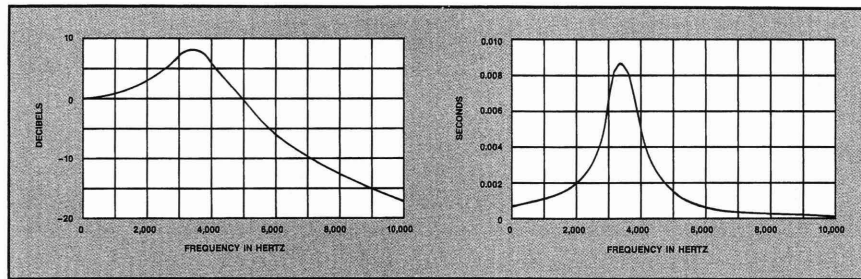


Figure 1. The audio magnitude and group delay of the transmit circuitry between the selected interface point and the VCO input.

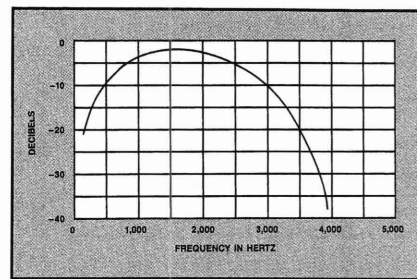


Figure 2. Audio spectrum of the 4,800bps Nyquist-filtered bipolar (NFB) signal.

With this filter, the cascaded response is $H_n(s) * H_d(s) / H_d(s) = H_n(s)$.

The pre-distorted eye pattern is shown in Photo 2 on page 34. When this signal is passed through the Mitrek transmit circuitry, the result is as shown in Photo 3 on page 34. Note how the resulting signal is symmetrical and how it converges to three distinct points.

Implementing pre-distortion

In the modem, digital transmit filters for many radio types are stored in memory. During initial setup, a filter is

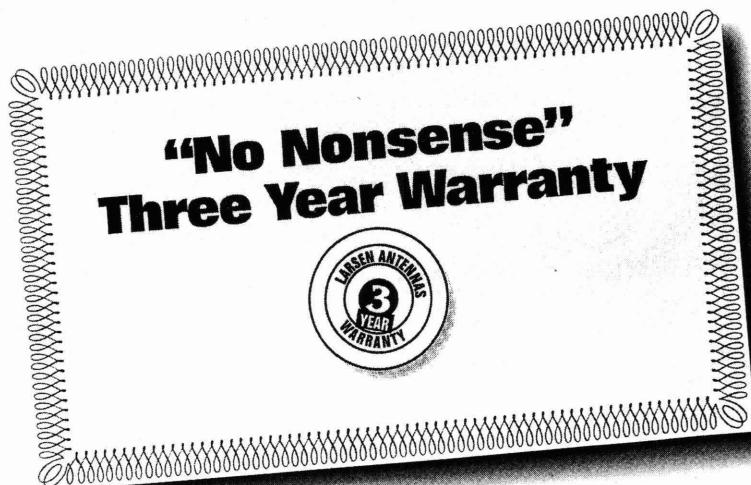
In the modem, digital transmit filters for many radio types are stored in memory.

selected that is matched to a factory-recommended transmit interface point.

Pre-distortion is a signal processing technique that allows high-speed data to be sent using voice-grade radio equipment—one of many uses for digital signal processing in mobile radio communications.

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Co-channel interference: How bad is it?

By Harold Kinley, CET

With the overcrowded land mobile radio frequency spectrum, interference has become the rule, rather than the exception.

Years ago, it was common for land mobile radio users to operate strictly carrier squelch. Although some users still get by with carrier squelch systems, their numbers are dwindling rapidly, and the time is fast approaching when it will be virtually impossible to operate without some kind of coded squelch system, digital or tone.

Classes of interference

Basically, interference can be clas-

sified as either of two types, *nuisance* or *destructive*.

Nuisance interference is heard in the absence of desired signals, but it does not significantly degrade the desired signal.

Destructive interference causes serious degradation to or even blocks the desired signal.

Although a coded squelch system does alleviate most of the *nuisance* interference problems, it is *not* the great panacea that it is perceived to be by most non-technical, and even some technical, people. It does nothing to alleviate problems caused by destructive types of interference.

Even nuisance interference can become harmful if allowed to go *untreated*. For example, a user who is constantly annoyed by noise or other undesirable signals often will turn down the radio's volume control or adjust the carrier squelch control to the maximum position, thus seriously desensitizing the receiver to even the desired signal transmissions.

At this point, the nuisance interfer-

ence has become harmful. A coded squelch system is the logical solution to this type of interference.

There is not much use in crying to the FCC about nuisance interference *unless* it is caused by an improperly or illegally operating transmitter that you can identify. You will not get much sympathy, especially if you are operating carrier squelch only.

Destructive interference is another matter.

Normally, great care is taken by the duly authorized frequency coordinators to eliminate or greatly reduce the likelihood of interference between two users operating on the same frequency.

In spite of the best efforts to avoid interference between co-channel users, problems often occur. Most of the

(continued on page 46)

Kinley is a certified electronics technician with the South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985.

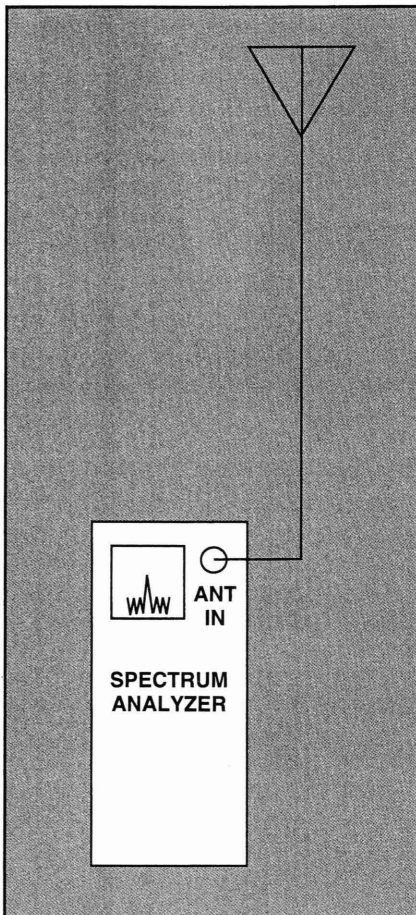


Figure 1. This simple setup can be used to measure the offending interfering signal's amplitude. The spectrum analyzer must be capable of *absolute* (dBm) measurement and not the simple *relative* dB indications on some imitation spectrum analyzers that are a part of some service monitors.

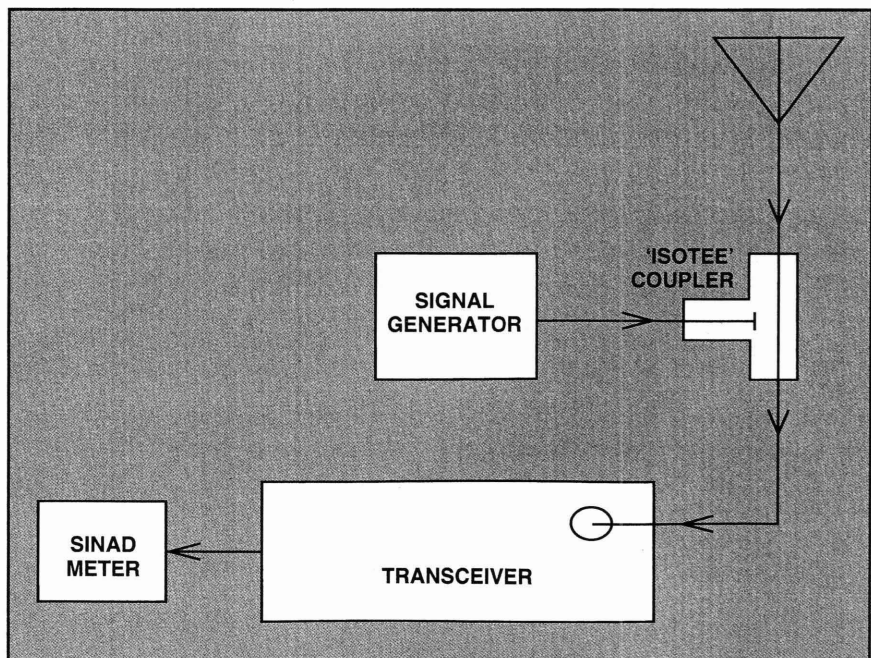


Figure 2. This setup can be used to measure the actual amount of degradation that the interfering signal causes to the desired signal. For example, suppose that in the absence of the interfering signal, a generator output level of -60dBm is required to produce the 12dB SINAD level. Also suppose that in the presence of the interfering signal, the generator output must be increased by 20dB to -40dBm to return the SINAD meter reading to 12dB SINAD . This means that the interfering signal is *degrading* the desired signal by 20dB .

Technically speaking

Formula 3

$\mu\text{V/m}$

$$= \text{antilog} \left[\frac{20\log(\mu\text{V}) - G_R + 20\log F + L_L - 32}{20} \right]$$

Formula 4

$\mu\text{V/m}$

$$= \text{antilog} \left[\frac{\text{dBm} - G_R + 20\log F + L_L + 75}{20} \right]$$

where

μV = signal level at receiver input in microvolts

dBm = signal level at receiver input in decibels above 1mW

G_R = antenna gain of the receiving antenna in dBd

L_L = line loss in decibels

F = frequency in MHz

(continued from page 8)

problems are of the nuisance type and can be solved by using different codes on coded squelch systems.

Compromising solutions

With destructive interference, a different approach is required. The best method of combatting co-channel destructive interference is for the two co-channel users to meet and to work out an arrangement that is mutually beneficial and acceptable to both parties.

Some of the factors that might be adjusted to minimize interference include the antenna height, antenna gain, antenna orientation, transmitter and antenna location, and transmitter power. It may be necessary to try a combination of several of these methods to effect a satisfactory solution.

Last resort

Obviously, if all else fails, it will be necessary to request intervention from the FCC.

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Such a request should be avoided if at all possible. If it is necessary, then you should be well prepared with documentation to show that the other party is indeed causing destructive interference to your radio system.

Logs showing dates and times of the destructive interference might prove helpful. Tape recordings of transmissions being blocked by the offending signal are useful in proving a case of blocking or serious degradation of the desired signal.

You should have plenty of recorded signal level readings on different dates and times for the FCC's examination and consideration.

Signal level measurements

The simplest method of measuring the offending signal's level is to use a spectrum analyzer as shown in Figure 1 on page 8.

Most spectrum analyzers measure signal level in dBm units. If you want to convert dBm units to microvolts, use the following formula:

$$\mu V = \text{antilog} \left[\frac{\text{dBm} + 107}{20} \right]$$

(1)

Example: Suppose, using the setup in Figure 1, the signal level was found to be -90dBm. Substituting into the formula, we have:

μV

$$= \text{antilog} \left[\frac{-90 + 107}{20} \right]$$

$$= \text{antilog} \left[\frac{17}{20} \right]$$

$$= \text{antilog} [0.85]$$

$$= 7.08 \mu V$$

(2)

This is the signal level at the spectrum analyzer input, and it would be the same at a receiver input.

Receiver method of measurement

Another way to determine the signal

level at the receiver input (when metering points are provided) is to check the meter reading at the IF metering point with the signal on the air.

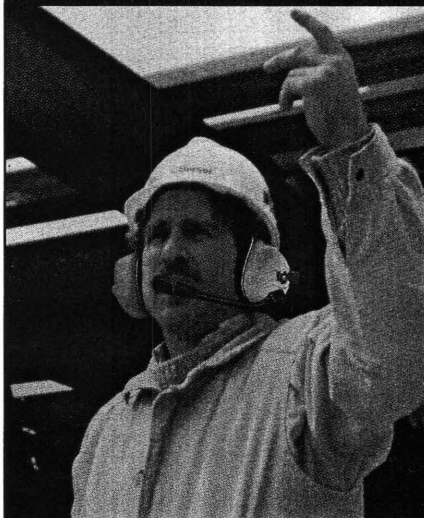
Record the meter reading for reference. Then, connect a signal generator to the receiver input and adjust the generator level to produce the same meter reading. Read the signal generator level in microvolts or dBm. This is

the signal level of the interfering signal.

Converting to $\mu V/m$

To convert the signal level at the receiver input in microvolts or dBm to field strength at the antenna in microvolts per meter ($\mu V/m$), use formula (3) or formula (4), as appropriate, in the box on page 46.

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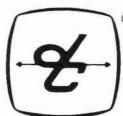
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Technically speaking

Example: The signal level in microvolts at the receiver input is $1.5\mu\text{V}$. The antenna gain is 6dB, and the transmission line loss is 3dB. The frequency is 160MHz. What is the field strength at the antenna in microvolts per meter?

Solution: Use Equation (3) and substitute:

Formula 5

$\mu\text{V/m}$

$$= \text{antilog} \left[\frac{20\log(1.5) - 6 + 20\log(160) + 3 - 32}{20} \right]$$

$$= \text{antilog} \left[\frac{3.52 - 6 + 44.08 + 3 - 32}{20} \right]$$

$$= \text{antilog} \left[\frac{12.6}{20} \right]$$

$$= \text{antilog} [0.63]$$

$$= 4.27\mu\text{V/m}$$

If the signal level at the receiver input were in dBm, you would use Equation (4) on page 46 to find the field strength in microvolts per meter at the antenna.

To determine how much the interfering signal is degrading the desired

signal, use the equipment setup shown in Figure 2 on page 8.

The *isotee* is a homemade coupling device that is very lossy and has a negligible loading effect on the line. It is made by removing the center pin from a T-connector, cutting it off short

and then reinserting it into the T-connector.

Then the female-female UHF adapter is attached to the T-connector. Because the T-connector's pin has been cut off, no direct contact is made between the double-female UHF adapter (barrel) and the T-connector; nevertheless, RF leakage within the connector provides some degree of coupling with considerable isolation between the signal generator and the line. Similar isolation couplers can be made with other types of RF connectors.

Measurement procedure

(1) With no incoming signal from the antenna and with the signal generator set to the correct frequency and modulated at $\pm 3\text{kHz}$ deviation, set the generator level to produce 12dB SINAD. Record the generator level in dBm. (It will take a high signal level to produce 12dB SINAD because of the loss or isolation of the isotee.)

(2) Wait for the interfering signal to make an appearance. When the interfering signal appears, increase the signal generator level to return the SINAD meter reading to 12dB SINAD. Record the new signal generator level.

(3) The difference between the generator levels in steps 1 and 2 in decibels is the amount of degradation the interfering signal is causing to the desired signal.

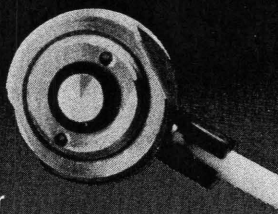
If your radio system operates at consistently stronger signals than the *minimum usable signal level* (12dB SINAD), then you might choose a higher input level as the reference. For example, $1\mu\text{V}$ might be a good choice if your signals are always this strong. Of course, the measurement would indicate less degradation at $1\mu\text{V}$ than it would at the $0.3\mu\text{V}$ level, for instance.

This information should help you to establish the magnitude or the severity of the interference problem. The severity of the interference may vary from day to day and time of day. If the problem occurs frequently and with severe degradation, it can render your radio system very unreliable.

But before taking your case to the FCC, arm yourself with information and documentation.

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Estimating radio range

By Harold Kinley, C.E.T.

Many variables go into the formula for calculating radio coverage area or range. Unlike free-space propagation, no simple formula fits every situation for plane-earth propagation. Of particular importance in estimating or calculating radio coverage is the terrain itself or the "lay of the land." In this column we will look at a series of range prediction graphs based on the modified Egli model—a formula found in Edward Singer's book, *Land Mobile Radio Systems* (Prentice-Hall, 1989).* This book is highly recommended for anyone work-

* This book is now in its second edition, published in 1994.

Kinley is a certified electronics technician with the South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985.

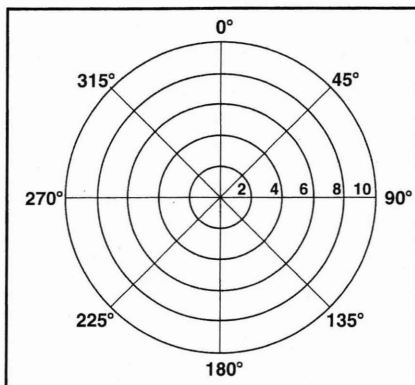


Figure 1. The intersections of cardinal radials from a tower site and concentric circles of a set increment are used as sampling locations for determining height above average terrain (HAAT).

ing in land mobile radio.

Before describing the use of these graphs, let's cover a few topics that you should know and understand to better use the graphs.

Effective base antenna height

To calculate the "effective" antenna

height or height above average terrain (HAAT) for a base antenna, three figures must be known or computed:

- (1) the height of the antenna above ground level (AGL).
- (2) the elevation of the tower site above sea level (ASL).
- (3) the average terrain elevation above sea level.

To calculate the average terrain elevation, you must know the elevation at the tower site and at 40 points surrounding the tower site. Figure 1 at the left shows the required locations where the elevation must be known. The center is the tower site location. Notice that there are eight radials extending from the center to a point 10 miles out. These radials begin at 0° and are spaced 45° apart. They are known as the *cardinal radials*. Concentric circles at two-mile increments are drawn around the center point, with the outer circle being at a distance of 10 miles from the tower or center. At each point where a radial intersects a circle, the elevation must be determined.

(continued on page 46)

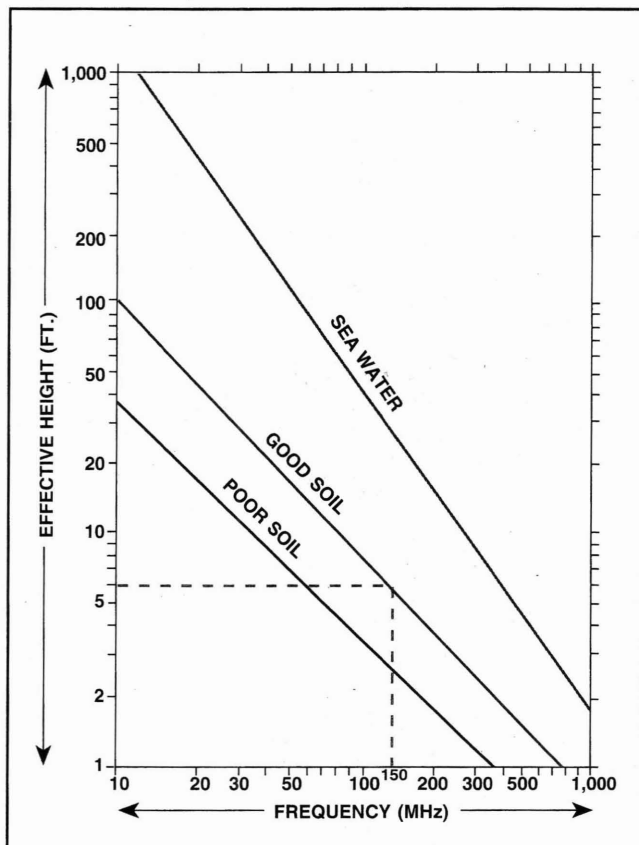


Figure 2. The effective height of a mobile antenna depends on the conductivity of the soil or other surface. Note that at 150MHz the minimum effective height is approximately 6 feet.

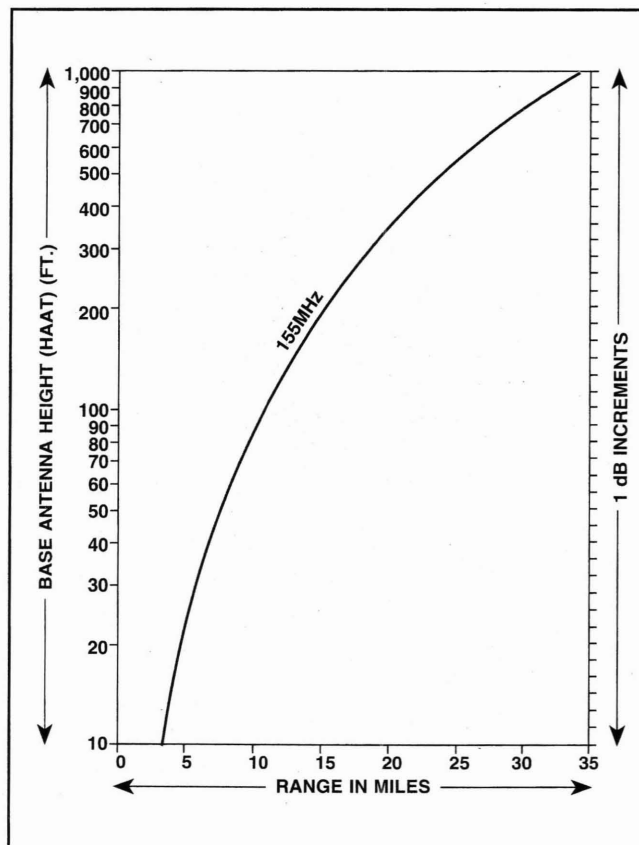


Figure 3. This is a graph of range vs. tower height for 155MHz with an effective radiated power (ERP) of 100W. The complete set of operating characteristics is: ERP = 100W; gain of mobile antenna = -1dBd; receiver line loss = 1.5dB; noise degradation = 3dB; receiver antenna height = 6 feet; probability of communication = 90%.

(continued from page 8)

A transparency of Figure 1 can be made to fit the scale of a topographic map and can be used as an overlay to determine the elevation of the various points required for calculating average terrain elevation. The tower site or center point is used eight times in the calculations, once for each radial, so it carries more weight than the other

points on the radials. Once the elevation at each of the points is determined, they are added together. Then the sum is divided by 48 to yield the *average terrain elevation*.

Once the average terrain elevation is determined, the HAAT, or effective antenna height, is determined by adding the antenna tower height above ground level (AGL) to the tower site elevation above sea level (ASL), then subtracting the av-

erage terrain elevation. The step-by-step process is shown in the table on page 48.

Effective mobile antenna height

For mobile antennas, a factor known as *minimum effective height* must be considered. The graph in Figure 2 on page 8 provides an approximation of the minimum effective mobile antenna height over *poor soil, good soil* and *sea water*. The dashed lines indicate that the minimum effective height of a mobile antenna at 150MHz is approximately 6 feet. The minimum effective height should be used for calculations if it is greater than the *actual* height of the mobile antenna. If the actual height of the mobile antenna is greater than the minimum effective height, then the actual height should be used.

Antenna gain figures

When using various formulas involving propagation and path loss, the *reference antenna* for the antenna gain figures must be clearly understood. A lossless halfwave dipole has a gain of 2.15dB over the isotropic radiator (a theoretical antenna with equal radiation in all directions). When the isotropic radiator is used as the reference, the gain should be stated in *dBi*. When the halfwave dipole is used as the reference, the gain should be stated in *dBd*.

Mobile antennas usually are referenced to the quarterwave mobile antenna. Compared to a halfwave dipole, the quarterwave mobile antenna would have a gain of about -1dBd.

Effective radiated power

The term effective radiated power (ERP) takes into consideration any losses or gains in the system between the transmitter output and the antenna (including antenna gain). It is easier to work with dBm or dBW units of measure because line losses and antenna gains are usually given in dB. To convert transmitter output to dBW, use the following formula:

$$\text{dBW} = 10\log P \quad [1]$$

To convert dBW back to power in watts use this formula:

$$\text{watts} = \text{antilog}(\text{dBW}/10) \quad [2]$$

As an example, suppose that a transmitter has an output of 100W, a transmission line loss of 2dB and an antenna gain of 5dBd. The transmitter output in dBW is:

$$\begin{aligned} \text{dBW} &= 10\log(100) \\ &= 10(2) \\ &= 20\text{dBW} \end{aligned}$$



Receive only	Freq. Ranges (MHz)	N.F. (dB)	Gain Comp. (dB)	1 dB Gain (dBm)	Device Type	Price
P30VD, P35VD, P40VD, P45VD	30-35, 35-40, 40-45, 45-50	<1.3	15	0	DGFET	\$ 44.95
P30VDG, P35VDG, P40VDG, P45VDG	30-35, 35-40, 40-45, 45-50	<0.5	26	+12	GaAsFET	\$109.95
P150VD, P160VD, P170VD	150-160, 160-170, 170-180	<1.5	15	0	DGFET	\$ 44.95
P150VDA, P160VDA, P170VDA	150-160, 160-170, 170-180	<1.1	15	0	DGFET	\$ 56.95
P150VDG, P160VDG, P170VDG	150-160, 160-170, 170-180	<0.5	24	+12	GaAsFET	\$109.95
P450VD, P460VD	450-460, 460-470	<1.8	15	-20	Bipolar	\$ 49.95
P450VDA, P460VDA	450-460, 460-470	<1.2	16	-20	Bipolar	\$ 74.95
P450VDG, P460VDG	450-460, 460-470	<0.5	16	+12	GaAsFET	\$109.95
P800VDG, P830VDG, P860VDG	800-830, 830-860, 860-890	<0.6	19	+12	GaAsFET	\$119.95

Inline (rf switched)

SP30VD, SP35VD, SP40VD, SP45VD	30-35, 35-40, 40-45, 45-50	<1.4	15	0	DGFET	\$ 74.95
SP30VDG, SP35VDG, SP40VDG, SP45VDG	30-35, 35-40, 40-45, 45-50	<0.55	26	+12	GaAsFET	\$139.95
SP150VD, SP160VD, SP170VD	150-160, 160-170, 170-180	<1.6	15	0	DGFET	\$ 74.95
SP150VDA, SP160VDA, SP170VDA	150-160, 160-170, 170-180	<1.2	15	0	DGFET	\$ 86.95
SP150VDG, SP160VDG, SP170VDG	150-160, 160-170, 170-180	<0.55	24	+12	GaAsFET	\$139.95
SP450VD, SP460VD	450-460, 460-470	<1.9	15	-20	Bipolar	\$ 79.95
SP450VDA, SP460VDA	450-460, 460-470	<1.3	16	-20	Bipolar	\$104.95
SP450VDG, SP460VDG	450-460, 460-470	<0.55	16	+12	GaAsFET	\$139.95

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The effective radiated power is:

$$20\text{dBW} - 2\text{dB} + 5\text{dB} = 23\text{dBW}$$

The ERP in watts is found from:

$$\begin{aligned}\text{watts} &= \text{antilog}(\text{dBW}/10) \\ &= \text{antilog}(23/10) \\ &= \text{antilog}(2.3) \\ &= 199.5\text{W}\end{aligned}$$

Noise degradation

If the mobile unit is operating in a heavily industrialized area, the noise level probably will cause significant degradation of the signal. That is, a higher signal level will be required to produce a given level of quieting or a given SINAD ratio from the receiver. The noise problem is less severe at VHF highband than on VHF lowband. As the operating frequency increases, the noise problem decreases.

According to *Electronic Communications Handbook* (1988 McGraw-Hill), the noise degradation in suburban residential New York is 10dB at 150MHz, 6dB at 450MHz and 0dB at 900MHz. In rural areas, the noise degradation would be much less: 2dB or 3dB or even less in extremely quiet areas. One noise factor is the ignition noise com-

RADIAL AZIMUTH	DISTANCE FROM TOWER IN MILES					
	0	2	4	6	8	10
0°	690	600	650	625	670	700
45°	690	570	590	615	650	625
90°	690	600	615	690	710	640
135°	690	625	640	675	700	680
180°	690	700	750	725	730	740
225°	690	680	680	690	640	650
270°	690	710	700	690	725	740
315°	690	710	740	750	710	700
TOTALS	5,520	5,195	5,365	5,460	5,535	5,475
	(A)	(B)	(C)	(D)	(E)	(F)
(1) Add column totals (A) through (F) = 32,550 (2) Average terrain = step (1) ÷ 48 = 678.125 (3) Tower site elevation = 690 (4) Antenna height = 500 (5) HAAT = (3) + (4) - (2) = 511.875						

Table 1. Step-by-step process for determining height above average terrain.

ing from the vehicle itself. This is one noise source that the mobile receiver cannot avoid. It can greatly degrade reception even at VHF highband frequencies.

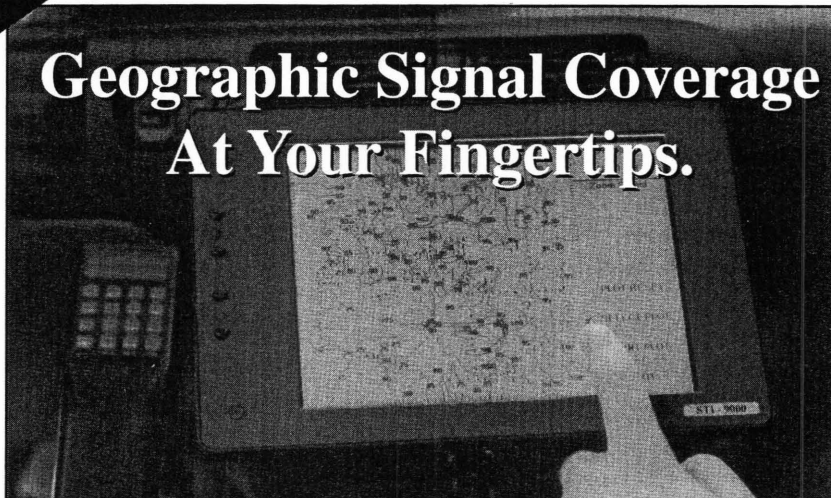
Reliability ratio

To improve the reliability or probability of communications, a fade margin or

reliability margin must be built into any formula for calculating radio coverage area or range. Higher probability factors require greater reliability ratios or fade margins in decibels. In the following graphs of radio communications range, the probability of communication is 90%. The reliability ratios are: 14dB for VHF

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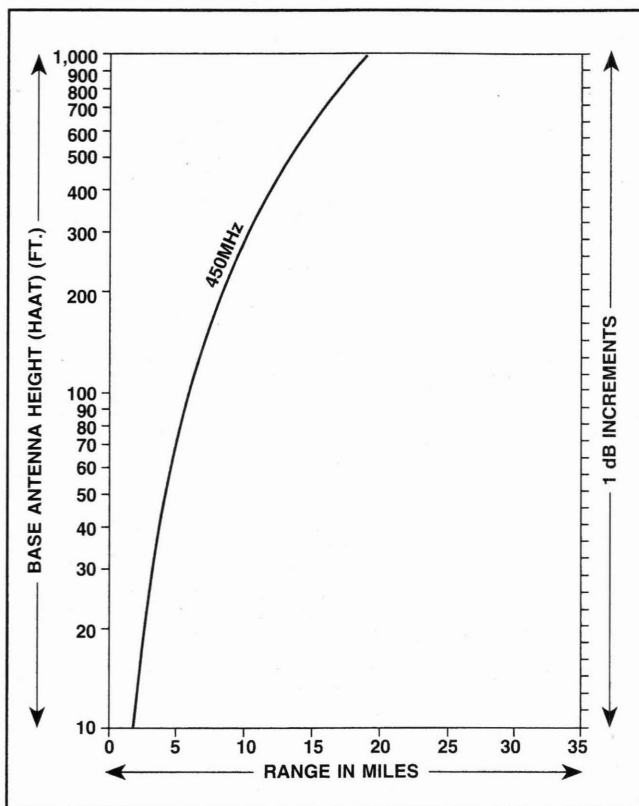


Figure 4. This is a graph of range vs. tower height for 450MHz with an effective radiated power (ERP) of 100W. The complete set of operating characteristics is: ERP = 100W; gain of mobile antenna = -1dBd; receiver line loss = 1.5dB; noise degradation = 1dB; receiver antenna height = 6 feet; probability of communication = 90%.

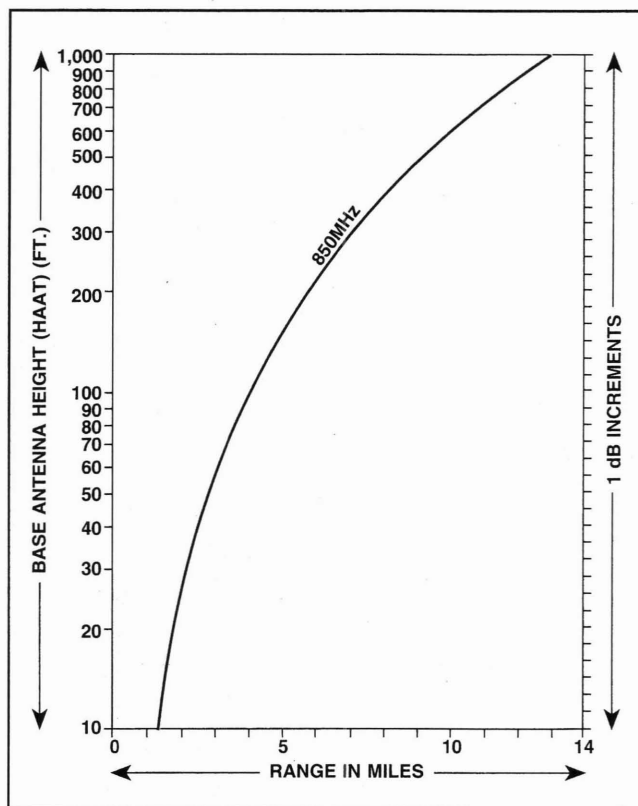


Figure 5. This is a graph of range vs. tower height for 850MHz with an effective radiated power (ERP) of 100W. The complete set of operating characteristics is: ERP = 100W; gain of mobile antenna = -1dBd; receiver line loss = 1.5dB; noise degradation = 0dB; receiver antenna height = 6 feet; probability of communication = 90%.

highband, 17dB for UHF and 19dB for 850MHz.

Radio coverage graphs

The graphs of radio range are based on terrain data from the Eastern Seaboard and Central Plains states with gently rolling ter-

rain and average hill heights of 50 feet.

Three graphs are presented: Figure 3 on page 8 for VHF highband (155MHz), Figure 4 above left for UHF (450MHz) and Figure 5 above right for 800MHz (850MHz). Because all of these graphs are similar, only one will be described here. The graph in

Figure 3 is for VHF highband. The various characteristics are listed. If the characteristics you need to use differ from those listed for this graph, adjustments can be made using the vertical decibel scale at the right. For example, this graph is based on an effective radiated power

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(ERP) of 100W. If the ERP you are using is different, simply apply the formula:

$$\text{dB} = 10\log(P/100) \quad [3]$$

where

P = effective radiated power in watts.

This calculation yields a dB correction figure to apply to the graph. For example, if the ERP is 200W, a +3dB correction figure results from Formula 3. Here is how to apply the correction figure to the graph.

Suppose, in Figure 3, that the antenna height (HAAT) is 200 feet. According to the graph, the range for 100W ERP is approximately 15.3 miles. If the effective radiated power is 200W instead of 100W, the dB correction figure will be +3dB. To get the range for 200W ERP, move from the 200-foot point over to the right dB scale and up 3dB. Move back over to the graph and down to the horizontal range scale to find the new range of approximately 18.3 miles for the corrected ERP figure.

Corrections for different noise levels also can be applied the same way. Receiver sensitivity also can be adjusted this way. For example, the graph in Figure 3 is for a receiver sensitivity of 0.35V (12dB SINAD). If your re-

ceiver sensitivity is 0.7V, the correction factor would be:

$$\begin{aligned} \text{dB} &= 20\log(0.35/S) \\ &= 20\log(0.35/0.7) \\ &= 20\log(0.5) \\ &= 20(-0.3) \\ &= -6\text{dB} \end{aligned}$$

where

S = 12dB SINAD sensitivity in V.

Be sure to observe the value signs of the dB correction factors. Negative correction factors reduce the range, and positive correction factors increase the range.

Figure 3 shows that doubling the antenna height is equivalent to a 6dB change. This relationship means that the same effect can be achieved by using a higher-gain base antenna, using a higher-gain mobile antenna, using better transmission line to reduce line losses, increasing transmitter power, using receivers with better sensitivity or any combination of the above that results in a net increase of 6dB.

Figure 3 also shows that a 6dB increase boosts the range by about 40%. A 12dB increase extends the range by about 100%.

For example, to increase the range by 40%, we could increase the antenna gain of the base station by 3dB and double the power (another 3dB increase).

Quadrupling the antenna height is equivalent to a 12dB change. This change is equal to increasing the transmitter power by a factor of 16. The increase in antenna height must be effective height or HAAT. For example, if the average terrain elevation is 750 feet, the antenna tower is 200 feet and the tower site elevation is 1,500 feet, then the HAAT is $1,500' + 200' - 750' = 950'$. If the actual tower height is doubled (from 200 feet to 400 feet) the new effective height is: $1,500' + 400' - 750' = 1,150'$. Thus, the effective height increased from 950 feet to 1,150 feet, much less than double. This change would result in only a slight improvement.

The graphs in Figures 4 and 5 are similar to the graph in Figure 3, but they are for 450MHz and 850MHz operations. Graphs such as these must be used cautiously and with two parts common sense! Even so, they are useful in making a determination of average coverages to be expected from given operating characteristics.

Computerized coverage plots

Computerized coverage plots have taken

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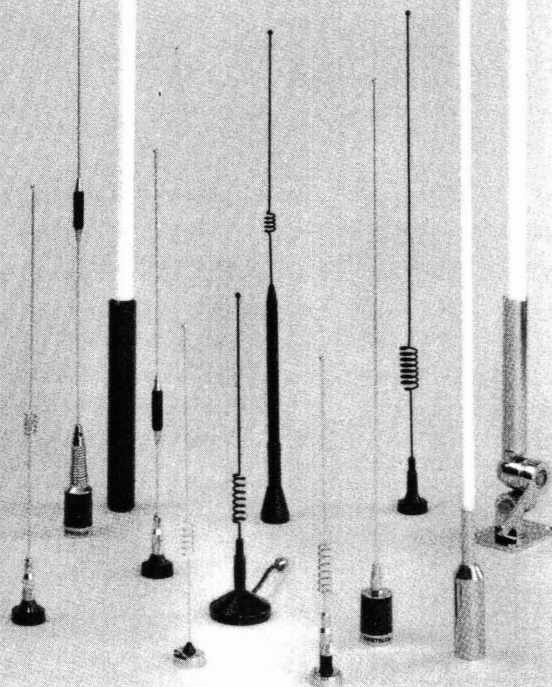
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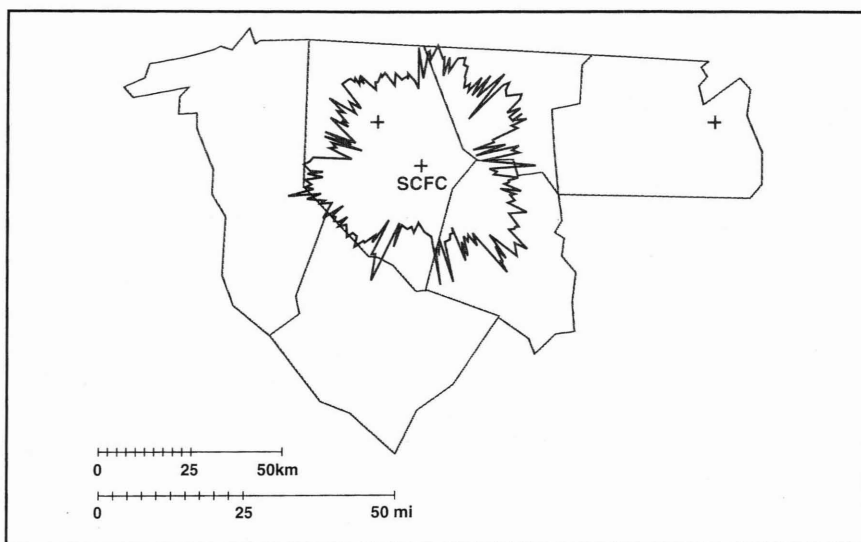


Figure 6. This radio coverage plot was produced with SoftWright's Terrain Analysis Package. The various terrain elevations are retrieved from a terrain database and the various elevations are computed for each point (at 0.1-mile increments) on each radial. This plot contained 360 radials. This contour is for 16dBu or 6.3 μ V/m. Considering a mobile antenna gain of -1dBd and a line loss of 1.5dB, this will produce slightly more than 1 μ V at the receiver input at 160MHz along the 16dBu contour.

much of the guesswork out of radio range calculation. Computer programs take into account effective radiated power, required signal level, probability of communications, ter-

rain elevations, antenna height and other pertinent factors. Coverage plots can be printed on a dot matrix printer or a plotter with county-line boundaries included on the plot.

Figure 6 at the left shows a computer-generated contour plot prepared with SoftWright's Terrain Analysis Package software. This plot was prepared using 360 radials at 1° spacing. The incremental value on each radial was 0.1 mile. This plot provided a contour with excellent resolution. The signal level for this contour was 16dBu. At 160MHz, with a line loss of 1.5dB and a quarterwave mobile antenna, the signal level at the receiver input is slightly higher than 1 μ V.

Several computer programs that are useful in determining signal level, radio range, path loss, HAAT calculations and ERP calculations are available from the author. The cost is \$10 plus \$2.50 for shipping and handling. Please specify 3½-inch or 5¼-inch floppy disk. These programs are for IBM-compatible PCs running MS-DOS. Write to the author at 204 Tanglewyld Drive, Spartanburg, SC 29301-2949.

Stay tuned!

References

1. *VHF and UHF Propagation*, Datafile Bulletin 10003-1, General Electric, Lynchburg, VA, 1962.
2. Inglis, Andrew F., *Electronic Communications Handbook*, McGraw-Hill, New York, 1988.
3. Singer, Edward, *Land Mobile Radio Systems*, Prentice-Hall, Englewood Cliffs, NJ, 1989.



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How to use line sweeping to maintain antenna systems

Two types of line sweeping equipment are suitable for feedline and antenna system measurements that detect deterioration and faults. Sample waveforms help to show you how to interpret test results.

By James M. Crooks

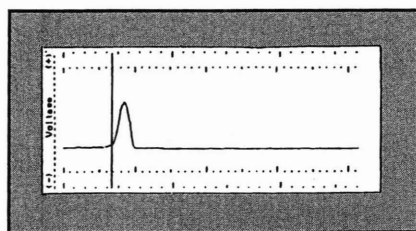


Figure 1. A properly **TERMINATED** cable absorbs the TDR signal, resulting in **no reflection**. Faults located ahead of the termination may be revealed as reflections along the waveform. If a terminator causes a reflection, the termination may be bad.

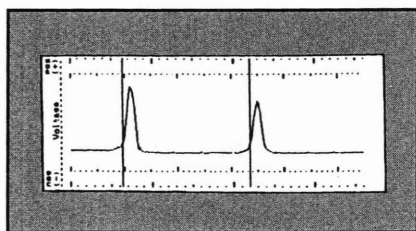


Figure 2. A reflection with the same polarity indicates a fault with **OPEN** (high impedance) tendencies. The reflection shown at the second cursor is a **COMPLETE OPEN**.

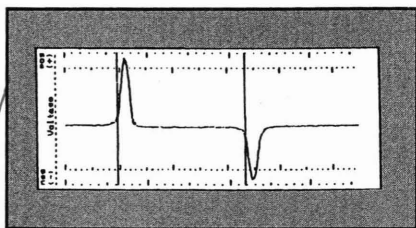


Figure 3. A reflection with the opposite polarity indicates a fault with **SHORT** (low impedance) tendencies. The reflection shown at the second cursor is a **DEAD SHORT**.

There appears to be a lot of confusion in the communications industry concerning transmission lines, antennas, connectors, waveguides and when and where to use time-domain reflectometers (TDRs) and frequency-domain reflectometers (FDRs).

The right information helps to dispel some of the misconceptions and helps technical supervisors to make important decisions about antenna system maintenance.

TDR

A TDR works on the same principal as radar: Pulses are transmitted down the cable under test and a part or most of the pulse energy is reflected.

The time required for the pulse to reflect and return to the source can be measured and plotted to provide a meaningful indication.

Time domain reflectometry is useful in three main areas of feedline and antenna system work:

- *Design*—It identifies impedance discontinuities.
- *Assembly and testing*—It aids in tuning.
- *Installation and maintenance*—It helps in fault location.

Crooks is president of Broadcast Communications, New Glarus, WI. He is a former chairman of the Wisconsin State 9-1-1 Board. He has been in the communications industry for 37 years, designing and installing systems for a wide variety of users.

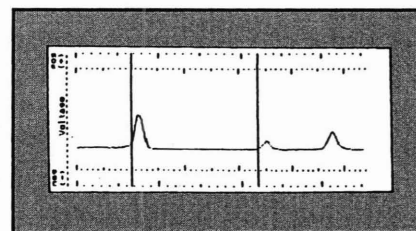


Figure 4. The middle reflection at the second cursor is a **PARTIAL OPEN** followed by a **COMPLETE OPEN** (end of the cable). The more severe the fault, the larger the reflection.

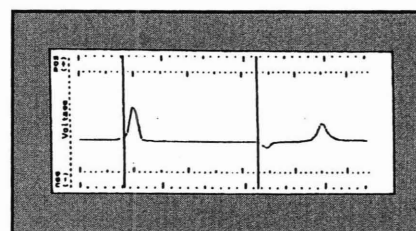


Figure 5. The middle reflection at the second cursor is a **PARTIAL SHORT** followed by a **COMPLETE OPEN** (end of the cable). The more severe the fault, the larger the reflection.

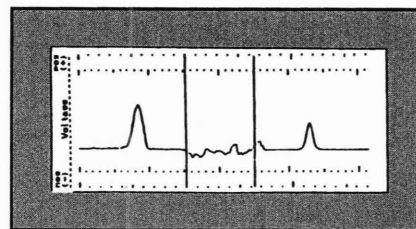


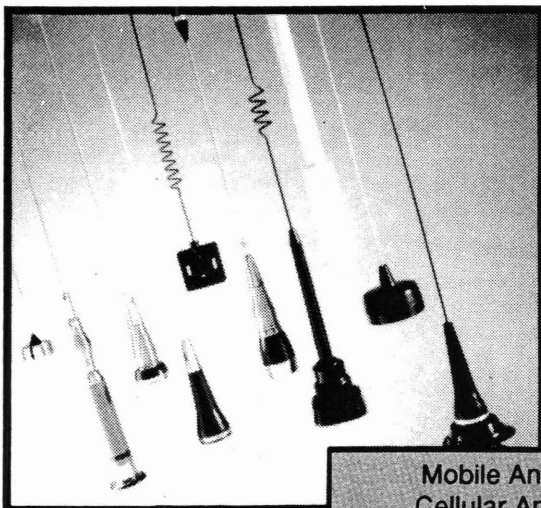
Figure 6. A water-soaked cable displays a waveform with a **downward slope**, indicating the beginning of the water, and an **upward rise** at the end of the water. Generally, the area in between the two reflections appears 'noisy'.



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Installation and maintenance is the most common use of a TDR.

Source pulse

The source pulse amplitude and duration determine its energy content and the maximum cable or line length that can be measured.

Pulse amplitude usually is fixed by the manufacturer and cannot be changed by the operator. Pulse width usually is adjustable via a preselection control that allows the operator flexibility.

The wider the pulse, the longer the feedline that can be accurately measured. There is a trade-off, though: A blind spot at the beginning of the pulse can mask discontinuities.

With a 2nS (2 nanoseconds or two billionths of a second) pulse, the blind spot covers approximately 3 feet; a

*The source pulse
amplitude and duration
determine its energy
content and the
maximum cable or line
length that can be
measured.*

10nS pulse has a 15-foot blind spot; and a 1,000nS pulse has a 600-foot blind spot. No return pulses are evident in the blind spot, so no faults are displayed within those sections of feedline.

One way to overcome the blind spot is to insert a jumper cable equal to the blind spot length between the TDR and the feedline to be measured. With a 2nS pulse, the jumper is 3 feet long. To calculate the distance to a fault or the feedline length, jumper cable length is subtracted from the TDR fault indication.

Reflected pulse

The reflected pulse amplitude always is less than the source pulse

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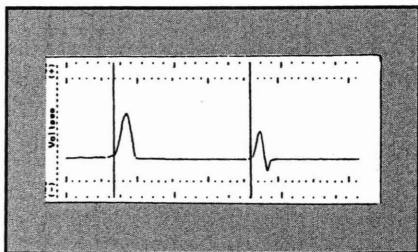


Figure 7. A HIGH RESISTANCE JOINT or SPLICE produces an *S-shaped* reflection. A high-impedance UPWARD reflection is followed by a lower-impedance DOWNWARD reflection.

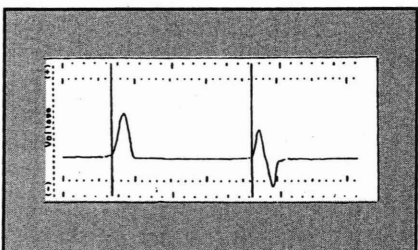


Figure 8. Testing through to an antenna usually results in an *S-shaped* reflection, although reflections can vary greatly depending on the antenna.

amplitude because of signal loss in the cable.

Keep in mind that the outgoing pulse is attenuated by line loss, and the return or reflected pulse is attenuated by the same amount.

Most TDRs present a positive-going source pulse, and the reflected pulse is either plus or minus, depending upon the return pulse's phase. Assuming a 50Ω source impedance, a positive return pulse indicates a reflected impedance greater than 50Ω (partial open or complete open of cable), and a negative return pulse indicates a reflected impedance less than 50Ω (kinks or water in the line or a partial short).

No reflection normally indicates a perfect 50Ω load that absorbs all of the pulse energy. An antenna connected to the line generally displays both a positive- and a negative-going pulse depending upon the antenna design and operating frequency.

TDR limits

Four important limitations should be

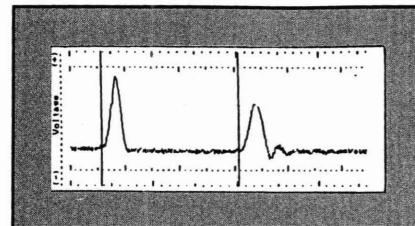


Figure 9. Testing tower cables with antennas can be challenging because of energy induction from high-RF areas as shown on the waveform. Stepping through various noise filter settings results in a cleaner waveform.

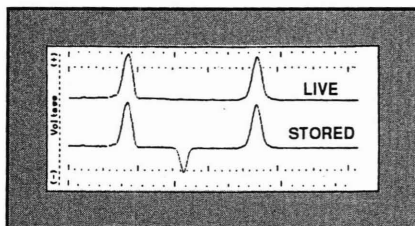


Figure 10. Some TDRs store images to allow an operator to compare the initial trace with another trace made after making a repair and print both traces as a permanent record.

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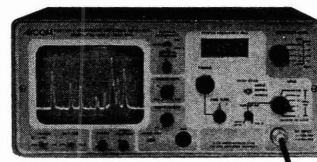
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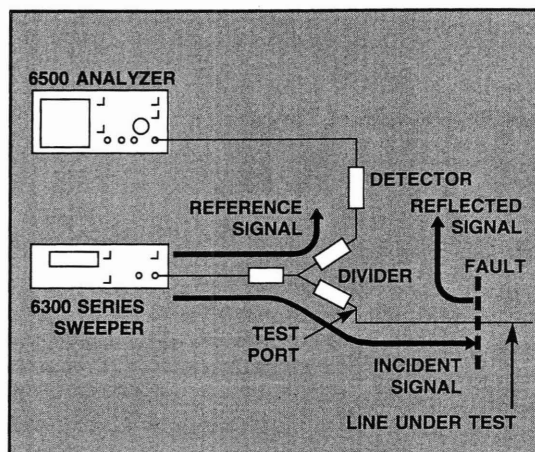


Figure 11. A scalar FDR setup block diagram for measuring distance to fault.

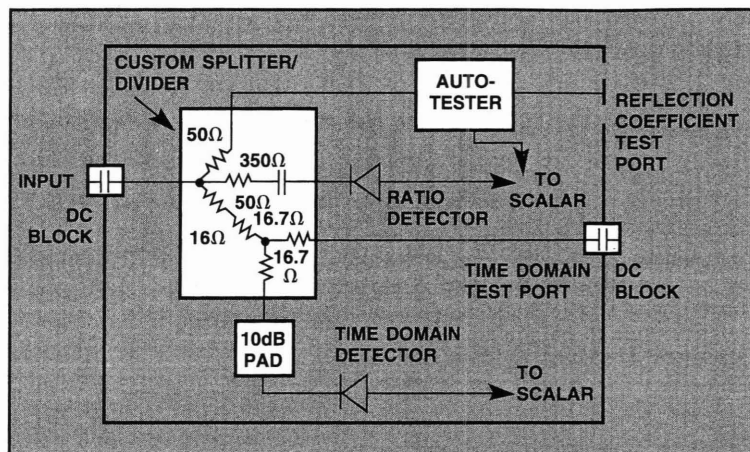


Figure 12. A scalar FDR setup block diagram for a directional coupler test head.

considered when using a TDR:

(1) The transmission line is pulsed at the TDR's pulse frequency, not at coaxial feedline's operating frequency.

Some line faults are frequency-sensitive and may not be detected by a TDR.

(2) An antenna cannot be checked for resonant frequency by a TDR.

(3) Waveguide cannot be measured by a TDR because of the waveguide's dispersion effects.

(4) The actual return loss of a coaxial line cannot accurately be measured by a TDR.

Although a TDR has limitations when used with radio communications systems, it provides sufficient informa-

tion in as much as 80% of transmission line applications, including:

- ▶ fault location.
- ▶ line length.
- ▶ types of faults.

Most TDRs are smaller than a briefcase. They usually are fitted with rechargeable batteries and a printer. The cost for most units ranges from \$2,000

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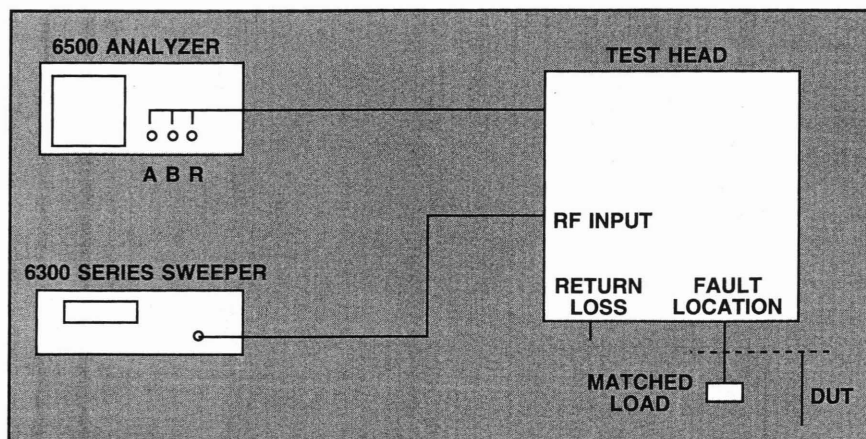


Figure 13. A scalar FDR setup block diagram for making insertion loss, return loss, VSWR, and antenna resonance measurements.

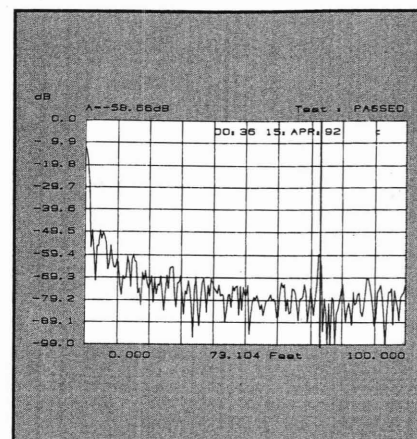


Figure 14. A typical plot of distance to fault of a transmission line.

to \$5,000.

The display can be as simple as a footage counter or as elaborate as a liquid crystal display (LCD) or oscilloscope image. Some TDRs store images to allow an operator to compare the initial trace with another trace made after making a repair and to print both traces as a permanent record. (See Figure 10 on page 32.)

Sample waveforms

Sample TDR waveforms show the operator what to expect on the display. For example, Figures 1-10 show:

- (1) a properly terminated cable.
- (2) an open cable.
- (3) a shorted cable.
- (4) a partial open.
- (5) a partial short.

A good approach is to obtain and read the TDR instruction manual to become familiar with the specific unit being used.

- (6) a water-soaked cable.
- (7) a high-resistance joint or splice.
- (8) an antenna connected to cable.
- (9) RF energy on cable (use filter).
- (10) dual trace, with both storage and live traces displayed.

The 10 waveforms show only a

small example of the many types an operator is likely to see.

Nevertheless, they represent the most common waveforms found in coaxial cable testing. The operator should become acquainted with the TDR and its various displays by experimenting with a variety of cables under simulated conditions. A good approach is to obtain and read the TDR instruction manual to become familiar with the specific unit being used.

FDR

A frequency-domain reflectometer works on the basis of a frequency-sweep generator: A sample of sweep generator signal is used to calibrate a scalar analyzer or tracking generator.

A directional coupler separates the

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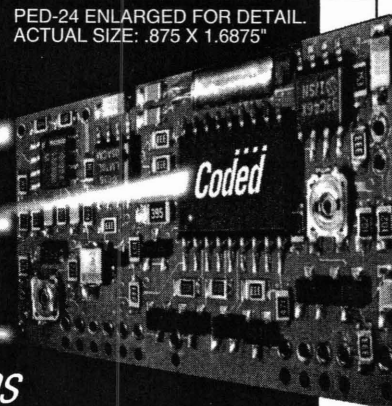
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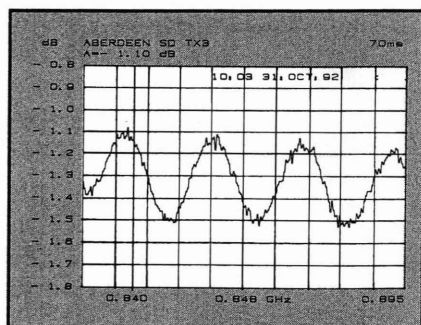


Figure 15. A typical plot of transmission line return loss.

outgoing sweep from the reflected sweep. A properly designed FDR with scalar data can provide the following information:

- (1) spectrum analysis.
- (2) insertion loss.
- (3) fault location.
- (4) antenna resonance.
- (5) return loss (voltage standing wave ratio or VSWR).
- (6) group delay.

The most commonly required measurements are spectrum analysis, insertion loss, fault location and antenna resonance.

► **Spectrum analysis**—A scalar FDR allows a frequency sweep of a transmission line, an antenna or both to display frequency-related problems in the line or antenna.

Antenna resonance is displayed to determine whether the antenna is defective or tuned to the proper frequency. A predetermined sweep range is selected, and any external signals being received by the antenna can be identified for receive level and frequency.

► **Insertion loss**—Insertion loss can be determined if both ends of a cable are available to measure the cable loss directly at the required operating frequency.

Loss in a jumper cable, connector, duplexer or combiner can be directly measured with the scalar FDR.

► **Fault location**—Although a TDR locates transmission line faults, it cannot measure the fault over a frequency spectrum, only time. (See Figure 11.)

A scalar FDR using the Fourier transform relationship can measure fault location at any predetermined frequency. This capability is extremely important at higher frequencies where

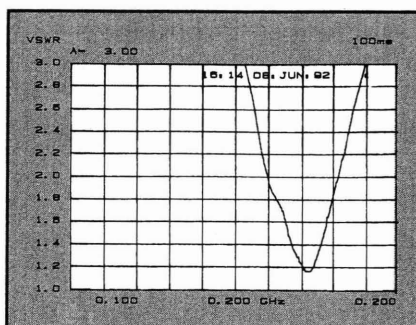


Figure 16. A transmission line VSWR plot.

cable connectors can affect VSWR greatly.

The FDR does not have a blind spot; the TDR is limited by the blind spot.

► **Antenna resonance**—The scalar FDR generates a curve of the frequency sweep and indicates the return

Good engineering practice dictates that repeating the sweeps every year helps to detect deterioration or increased return loss.

loss or VSWR at all sweep frequencies; therefore, the antenna's resonant frequency is plotted, as well as the return loss.

► **Return loss**—Perhaps one of the most important and popular measurements, *return loss*, allows the operator to measure the amount of line loss

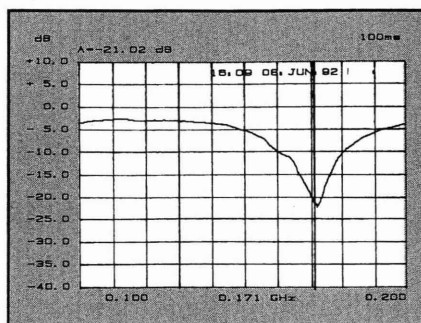


Figure 17. An antenna resonance plot.

directly at any operating frequency.

This loss can be checked against the calculated loss provided by the cable manufacturer. This test determines the cable's condition and measures return loss accurately. The operator must know the operating frequency, cable length and cable velocity of propagation (VOP).

From the amount of return loss, the operator can calculate or directly read the VSWR or the *voltage reflection coefficient* of the cable. The smaller the VSWR reading, the better the cable and load impedances match.

Figures 12-18 show the block diagram of the directional coupler test head for a scalar FDR; the block diagram of the scalar FDR setup for making insertion loss, return loss, VSWR and antenna resonance measurements; and several typical test plots.

Other information

Typically, TDR measurements are used for frequencies as high as 450MHz on coaxial cables, which covers as much as 80% of the applications.

Above 800MHz and at microwave frequencies, the FDR is highly preferred.

A scalar FDR costs \$40,000 to \$60,000, which is a substantial investment for a smaller company. Most companies contract a line sweeping service to check feedlines when they are installed.

Good engineering practice dictates that repeating the sweeps every year helps to detect deterioration or increased return loss. Early detection allows repairs and replacements to be scheduled before a serious loss occurs.

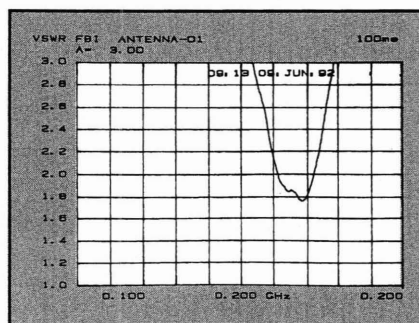


Figure 18. An antenna VSWR plot.

Digital technology serves 2-way radio communications

Linear predictive vocoding digitizes voice on radio communications channels. How users migrate to digital technology has an important bearing on choosing among the three access methods.

By Tim Twohig
and Vicki Griesser

Digital voice modulation may relieve land mobile radio channel congestion, increase data transmission capacity and add advanced features.

Digital enhancements

► **Audio quality**—New linear predictive voice encoding (vocoding) techniques and forward error correction improve digital audio quality.

Semiconductor technology advances have produced digital signal processors (DSPs) that execute millions of instruc-

tions per second. Linear predictive coders use the DSPs' processing power to convert the analog voice signal to a digital signal using computational approximation. This processing type provides digital audio quality comparable to analog.

Digital systems can provide uniform audio quality throughout more of a system's coverage area. Near the transmitter, digital and analog audio quality are comparable. At the coverage area fringe, where analog audio quality suffers from noise and fading, digital audio with forward error correction can be much better. (See Figure 1 below left.)

As the radio signal weakens, noise and fading cause digital bitstream errors. The digital receiver compares inbound information with codes that correct bitstream errors to reconstruct the

transmitted audio (or data message) accurately.

► **Data capability**—Digital technology can increase a radio system's data capabilities in a number of ways.

On a digital system, voice and data are transmitted in digital form, making it easier to integrate them. Data and voice transmissions can be carried on a single channel without one interfering with the other.

Digital modulation schemes support higher data transmission rates than today's two-state frequency-shift keying (FSK) modulation. Some of the next-generation digital systems will use multistate modulation techniques such as quadrature phase-shift keying (QPSK) and quadrature amplitude modulation (QAM).

These modulation techniques allow

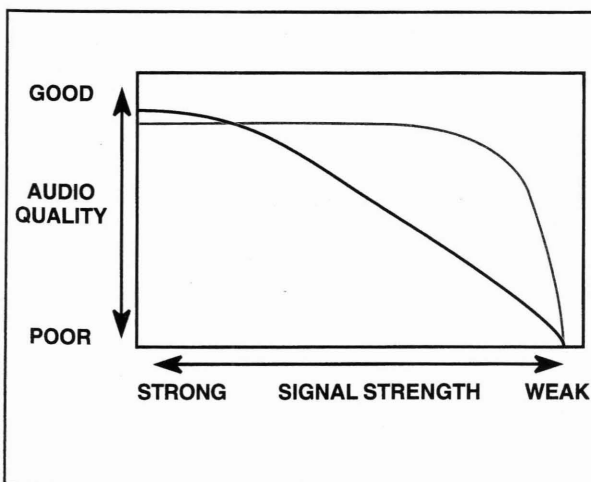


Figure 1. Digital audio quality surpasses analog quality as the radio signal weakens. At the fringe of the coverage area where analog audio quality suffers from noise and fading, digital audio with forward error correction can be much better.

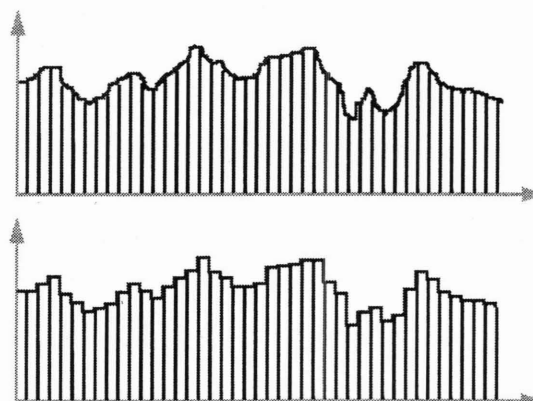


Figure 2. Pulse code modulation (PCM) samples the analog voice signal many times per second. Each sample is used to reconstruct the analog signal. At a rate that maintains high voice quality, PCM sampling generates more data than can be sent over a 25kHz-wide radio communications channel.

Twohig and Griesser are systems engineers at Motorola, Schaumburg, IL.

higher data rates by representing multiple bits with each state change. Using higher data rates makes applications that are too data-intensive for two-state modulation more practical. Some applications include data transmissions to and from laptop computers, peripherals and facsimile machines.

► *Spectrum efficiency*—Digital technology can carry more communications

on a given amount of radio spectrum, improving spectrum efficiency.

Voice modulation evolution

Long-distance telephone carriers were the first companies to digitize voice in the 1960s using 64kbps pulse code modulation (PCM).

PCM samples the analog voice signal level 8,000 times per second. Each

sample is represented by 8 bits of digital information that are used to reconstruct the analog signal. (See Figure 2 on page 11.) This coding method provides voice transmissions over long-haul wires with minimum noise and distortion.

When digital voice is sent over radio, the amount of data that can be sent over a given bandwidth is limited to provide acceptable performance to adjacent-channel users. With two-state digital modulation, it is impossible to transmit 64kbps digital signals over a 25kHz radio channel.

In the 1970s, vocoding techniques such as continuously variable slope delta (CVSD) modulation accurately represented a human voice using only 12kbps of information. This amount of data fits on a 25kHz channel with two-state digital modulation.

Unlike PCM, which samples the analog signal's level, CVSD samples its slope. The digital signal, made of binary (ones and zeros) codes, approximates the rising slope values with ones and falling slope values with zeros. (See Figure 3 below.)

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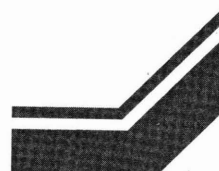
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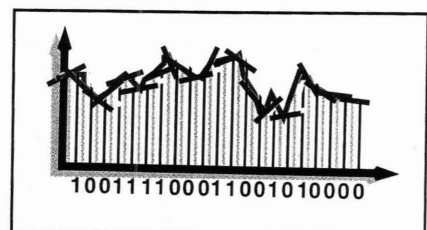


Figure 3. Continuously variable slope delta (CVSD) modulation samples the analog signal's slope. The digital signal, composed of binary (ones and zeros) codes, approximates the rising slope values with ones and falling slope values with zeros.

One bit of information represents each sample. Because CVSD vocoders degrade audio quality slightly, they are used only in encryption systems.

PCM and CVSD convert analog voice to digital voice by approximating the analog signal. Linear predictive vocoding uses the DSP's computing power to digitize voice by approximating the voice process.

The physical process of producing speech can be viewed as the mouth, nasal cavities and tongue acting to shape air impulses produced by the vocal cords (pitch is the frequency of these im-

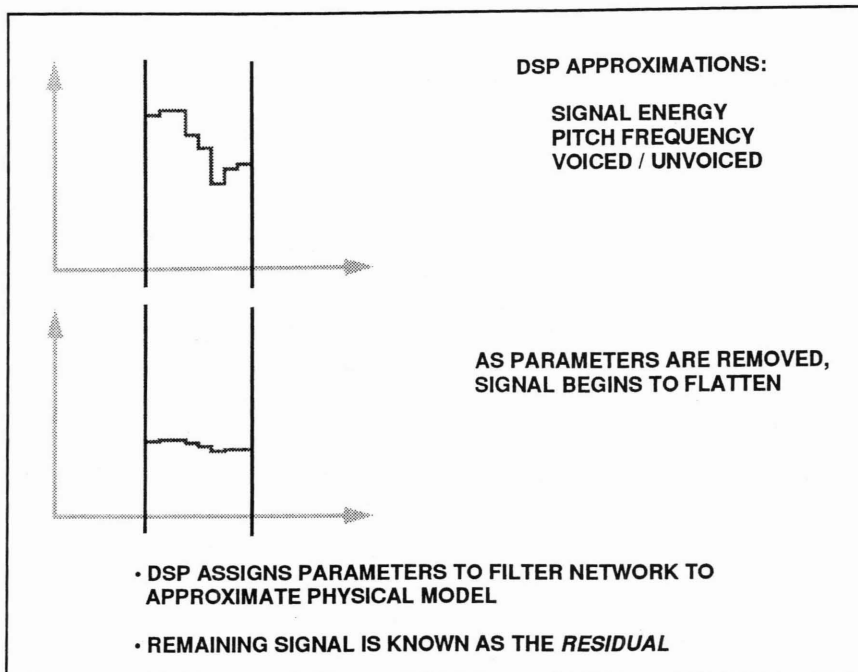


Figure 4. Linear predictive coding transmits filter coefficients that approximate speech patterns. Additional speech pattern information stored in the receiver is accessed by residual sequence address codes and combined with the filter coefficients to improve the digital audio quality.

pulses) for voiced sounds or air noise from the lungs for unvoiced sounds. Linear predictive coding derives coefficients for a filter structure that approximates this process.

The linear predictive coding process has several steps. (See Figure 4 to the left.)

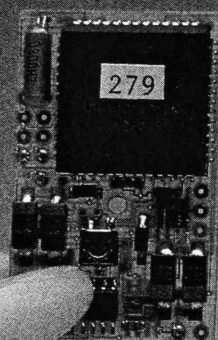
First, the analog signal is digitized using 64kbps PCM. This digital stream is divided into relatively short speech segments that the DSP processes. The DSP derives the filter coefficients that approximate the speech pattern and sends them to the transmitter.

Reconstructing the signal from these coefficients alone results in understandable, yet mechanical sounding, speech.

Audio quality improvements are achieved when the *residual* sequence, which is the difference between the input signal and the approximated signal, is transmitted. But transmitting the entire residual sequence results in a higher bit rate requirement, which in turn affects this method's efficiency.

To solve this problem, a table of

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residual sequences was derived. Instead of sending the entire residual sequence, only the residual sequence's address is transmitted. The receiving radio looks up the residual sequence in its electronic memory and adds it to the approximated signal.

Linear predictive coding requires approximately 4.8kbps of information to provide audio quality comparable to

analog. This information, when combined with new modulation techniques, easily fits a 25kHz channel.

In fact, there is much more room than what is needed to send just one voice over the channel. The additional capacity makes digital radio systems more spectrally efficient than their analog counterparts.

Three digital communications tech-

nologies, FDMA, TDMA and CDMA, are suited to different radio communications applications uniquely. The technologies differ in the way they make use of the additional capacity.

► **FDMA**—Frequency division multiple access (FDMA) offers more capacity than necessary to send one 4.8kbps digital voice over a 25kHz channel, so a smaller channel or a channel with less bandwidth may be used.

One implementation of FDMA technology uses a 12.5kHz-wide channel (half the bandwidth of previous channels) and provides the capability to create twice as many channels as 25kHz-wide channels from the same amount of RF spectrum. (See Figure 5 below.)

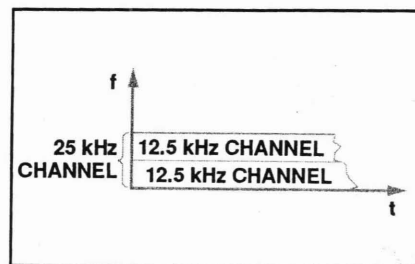


Figure 5. One implementation of FDMA technology uses a 12.5kHz-wide channel (half the bandwidth of previous channels) and provides the capability to create twice as many channels as 25kHz-wide channels from the same amount of RF spectrum.

Results of the FCC *Notice of Inquiry* concerning spectrum efficiency in land mobile frequency bands show that nearly two-thirds of respondents favor narrowband operation to reduce frequency congestion. The remaining respondents had no negative comments regarding narrowband FDMA.

There are two methods of converting existing wideband channels to digital narrowband channels. (See Figure 6 on page 18.)

The *on-channel* method places narrowband digital channels on the existing 25kHz center frequency. When adjacent 25kHz channels have been converted to narrowband channels, the 12.5kHz space between them can be occupied by a new narrowband channel.

The *offset* method places two narrowband channels in the space formerly occupied by one wideband channel. The center frequencies of the two new channels are offset 6.25kHz to either side of the wideband channel center.

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Any benefit realized through channel-splitting or converting is contingent upon local and national regulatory guidelines.

FDMA techniques have been optimized for use in the land mobile environment. A 12.5kHz digital channel using QPSK-C modulation can transmit 9.6kbps of data while maintaining the required level of adjacent channel protection.

Site issues, such as transmitter noise and receiver desensitization, still must be evaluated (as they would be with analog technology).

Because only 4.8kbps is required to represent a voice, extra capacity is available.

The extra capacity can be used to send other information such as error correction codes to ensure high audio quality; system signaling to add new features and improve existing features; and encryption information to ensure secure communication.

In today's analog systems, voice and data contend for a channel. The signal-

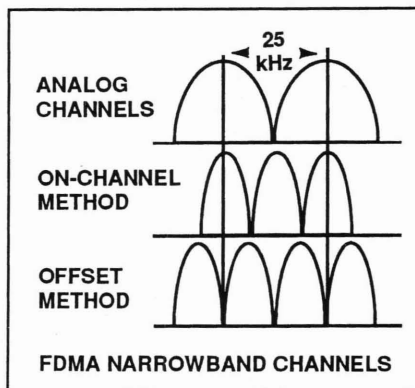


Figure 6. Converting 25kHz analog channels to 12.5kHz digital channels may be accomplished via the on-channel or offset method.

ing information is sent first, and the voice message follows. This process truncates the voice.

The digital system uses a process called *embedded signaling* that intermixes the extra 4.8kbps of signaling information with the digital voice.

Embedded signaling eliminates truncation and adds reliability into the sys-

tem by embedding system and radio information continuously through the duration of the voice message. Features such as encryption, emergency, unit ID, automatic power control and alphanumeric messaging may be realized and made more reliable by embedding the signaling in the voice message.

Data capacity increases significantly when the communication is entirely in a data mode. Field units operating in the data mode send as much as 7.2kbps of data and 2.4kbps of error correction. Not only can digital technology provide a dramatic increase in data capacity compared to today's analog systems, but the variety of data applications can be increased as well.

Error correction codes help to ensure data integrity, adding reliability.

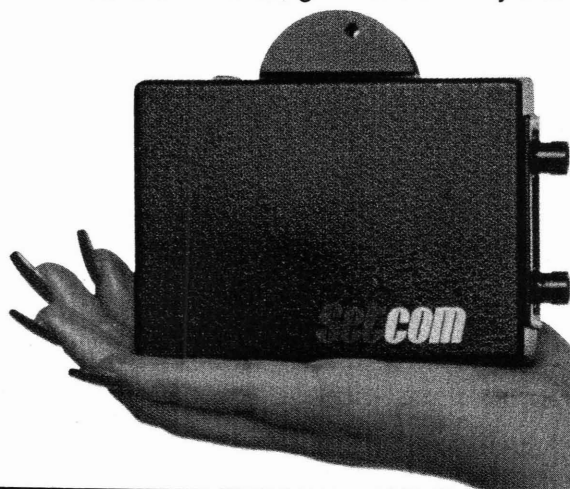
FDMA systems can increase spectrum efficiency by 2:1, add new features, enhance existing features and increase data capacity along with providing high audio quality in trunked and conventional radio communications systems. FDMA systems can be used in any fre-

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quency band.

► **TDMA**—Time division multiple access (TDMA) uses a 25kHz channel's additional capacity to transmit multiple voice messages simultaneously. Spectrum efficiency is accomplished by splitting the channel into repeating time slots, each of which can carry the necessary information for one voice conversation. (See Figure 7 to the right.)

One TDMA access method splits the 25kHz channel into six repeating time slots, potentially enabling as many as six users or groups of users to share the channel at any time.

TDMA technology inherently provides multiple communications paths. To use these paths effectively, some control must be placed over path access and allocation to user groups. Trunked ra-

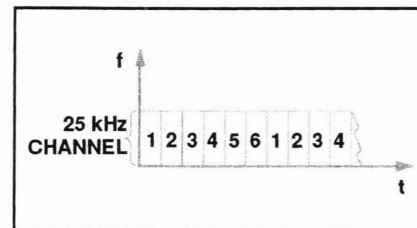
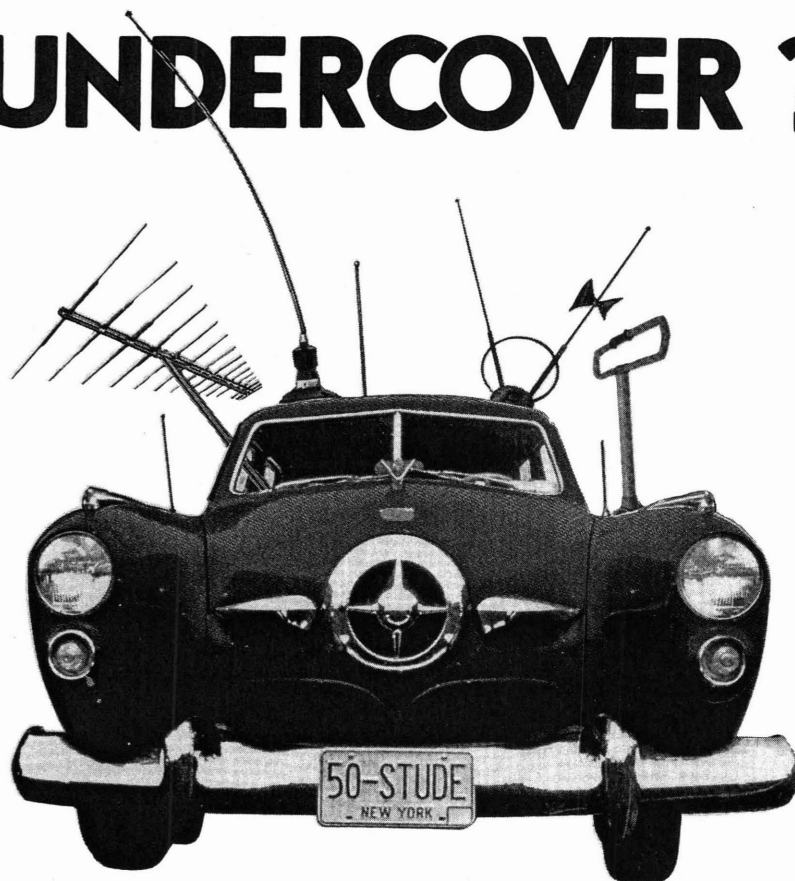


Figure 7. Time division multiple access (TDMA) uses a 25kHz channel's additional capacity to transmit multiple voice messages simultaneously. Spectrum efficiency is accomplished by splitting the channel into repeating time slots, each of which can carry the necessary information for one voice conversation.

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dio technology provides the control and dynamically assigns these talk paths to users upon request.

The TDMA modulation technique has been maximized for the land mobile environment, allowing digital channels to coexist with adjacent analog channels. Nonetheless, co-channel use is a spectrum issue that must be addressed.

The mobiles and portables in this type of system must be synchronized tightly for orderly talk path assignments. Any interference that causes errors in the outbound signaling could jeopardize this synchronization. Therefore, exclusive-use channels, such as those available at 800MHz, are strongly recommended for any TDMA communications system.

Because each user only transmits and receives one-sixth of the time, the transmitter and receiver each can be assigned a different time slot, allowing full-duplex telephone interconnect calls on mobiles and portables without the use of a duplexer. With TDMA, full-duplex interconnect becomes an inherent capability, resulting in added convenience and improved economy in telephone communications.

Systems using TDMA can support data and voice transmissions on one channel using different time slots.

For example, one individual could use one time slot for a voice conversation while another individual faxes documents on another time slot.

A new linear modulation technique, M-16QAM, transmits information at 64kbps over a 25kHz channel. The higher data rates allow data-intensive applications, such as facsimile, on a TDMA system.

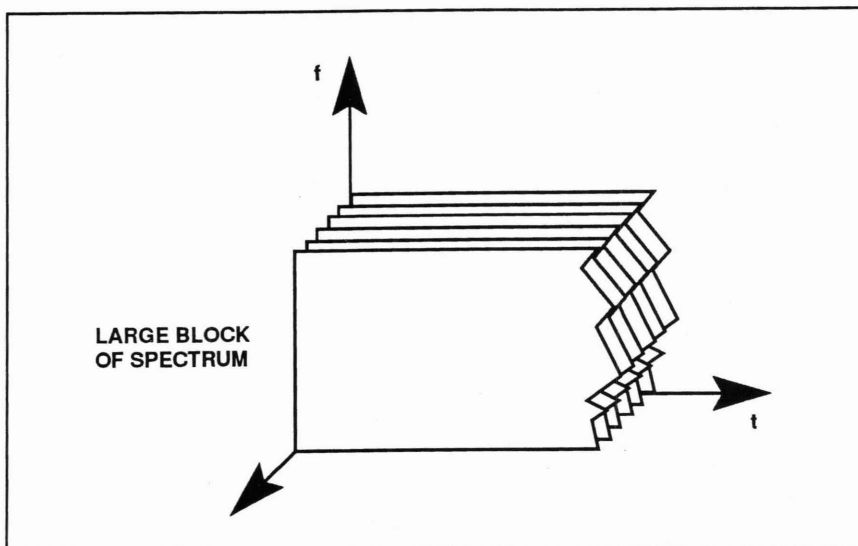


Figure 8. Code division multiple access (CDMA or digital spread spectrum) spreads the transmitted energy over a large amount of spectrum, using a low-powered signal that resists interference. Each transmission is coded so that CDMA receivers can separate the messages they are intended to receive.

Because of its dramatic increase in spectrum efficiency, TDMA technology is particularly useful in large

metropolitan areas where the radio spectrum is congested. TDMA's 6:1 increase in spectrum efficiency can help

to relieve overcrowding on some channels.

The technology will be useful in systems with many users sharing one RF backbone, such as specialized mobile radio (SMR) systems.

TDMA digital systems will offer full-duplex telephone interconnection, increased data capacity and high audio quality to improve trunked radio communications.

► **CDMA**—Code division multiple access (CDMA) technology uses even wider bandwidths than existing 25kHz channels to gain capacity. (See Figure 8 to the left.)

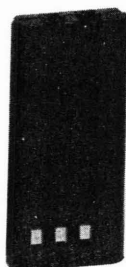
Also known as *digital spread spectrum*, CDMA spreads the transmitted energy over a large amount of spectrum, using a low-powered signal that resists interference. Each transmission is coded so that CDMA receivers can separate the messages they are intended to receive.

The combination of low-power signal and coding could allow many users to share the spectrum without interfering

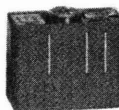
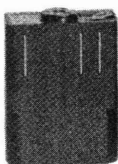
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with one another. Nevertheless, each spread spectrum user generates some degree of interference to other users, which limits the number of users who can share a given amount of spectrum.

High increases in data rates could be possible due to the wider channel bandwidths. With CDMA's extremely wide bandwidth, even more data-intensive applications could be made possible. Fu-

ture CDMA implementation could make video transmission over radio communications channels possible.

CDMA's inherent low power requires many base station sites to cover even a relatively small geographic area. Coordinating a large number of sites for multi-radio group calls could bog down even the most advanced audio switches available today.

CDMA's low-power transmissions would have problems coexisting with present radio systems. Overlaying CDMA on existing spectrum would raise the noise floor for the existing users, resulting in reduced coverage and interference for those existing users.

Frequency coordination would be necessary to implement CDMA in land mobile radio to avoid interfering with existing users.

Migration

Although digital technology is a significant step in two-way communications, many users' requirements will continue to be met with analog technology for years to come.

Manufacturers must consider several migration alternatives to address the wide variety of existing analog systems. The goal is to trade out the analog system as required and maintain interoperability between analog and digital radios.

To complete a migration, users eventually will need to incorporate digital radio equipment in mobile and portable units, as well as a fixed "channel" infrastructure.

A migration must be equally flexible for single-site systems and large multi-regional systems. If a migration path is laid out properly, the analog-to-digital transition would add digital benefits where needed without disrupting an existing user's radio activities.

Summary

FDMA, TDMA and CDMA can offer land mobile radio significant improvements over today's analog systems.

Each of these technologies has unique applications. FDMA provides advanced new features and improves existing features, and it improves spectrum efficiency.

TDMA provides a dramatic improvement in spectrum efficiency, along with full-duplex telephone interconnect on all mobiles and portables.

CDMA has tremendous potential for improvements in features and spectrum efficiency, but still is in early stages of research.

An important element of any digital strategy is a migration plan. A well-planned migration allows users to implement enhancements provided by digital technology.



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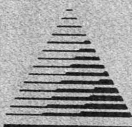
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Designing microwave radio links

They remain the preferred backhaul choice for public safety radio networks

The golden age of point-to-point microwave was the 1960s, when nearly all long-distance telephone calls and all network television signals traveled over the medium. Eventually, fiber optic cable replaced microwave for most telephone calls, and satellite took over for television.

Today, cellular radio networks rely primarily on landline T-1s for backhaul, but most public safety radio networks use microwave radio for this purpose. Microwave is preferred by public safety because the recurring costs are lower than leased T-1s, and reliability can be higher than landline networks. As an old AT&T microwave engineer once said to me, a landline can be cut at an infinite number of points, but a microwave link can fail at only two spots. Modern microwave links often are organized into self-healing rings with equipment and path redundancy to ensure a single equipment failure does not cause a network failure.

Microwave paths are quite different from land mobile radio paths. Unlike mobile radio, microwave radios are always stationary, and every effort is made to ensure the antennas are line-of-sight. The dominant impairment should be free space loss, and fading occurs not from scattering but from atmospheric effects.

Microwave paths must have an extremely high availability (greater than 99.995%) and low bit-error rate (less than 10^{-9}) in order to be effective. Bandwidths of microwave channels are much wider than most mobile radio channels, usually on the order of 30 MHz, with bit rates as high as 155 Mb/s. To achieve these bit rates, higher-order signal constellations such as 64-QAM are used with trellis-coded modulation.

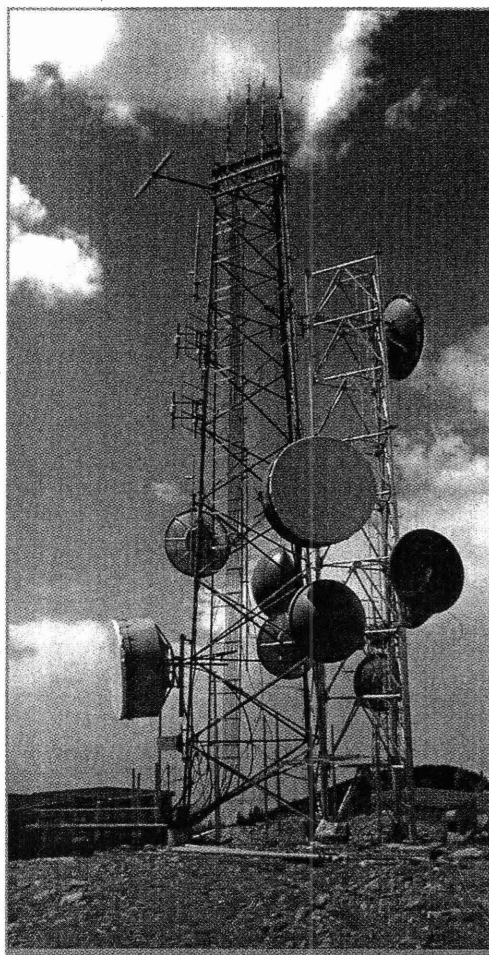
In the U.S., the FCC allocates spectrum for microwave point-to-point

links, and frequency coordination is required as part of the license application. Antenna gain and polarization are used to pack frequencies very tightly. Depending on the band, high-performance antennas with reduced sidelobes may be required to mitigate interference to other systems. Popular microwave bands include the 2, 4, 4.9, 6, 10, 12, 18 and 23 GHz licensed bands and the 2.4 and 5.8 GHz unlicensed bands.

This article describes the process for designing microwave links, which involves picking the sites, plotting the path profile, predicting the path loss and calculating the path availability.

Site selection. In LMR networks, microwave radio sites are usually co-located with the LMR tower and the tower height is used to ensure a line-of-sight path. Sometimes an intermediate relay is necessary when terrain or buildings block the path.

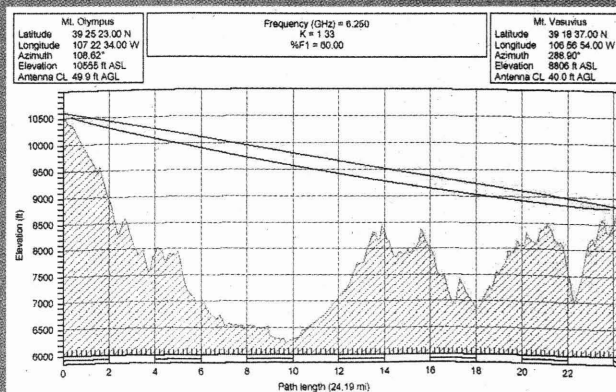
Path profile. After the prospective sites are chosen, the



Microwave and LMR antennas on Tesuque Peak in New Mexico.

Figure 1

Typical unobstructed path profile with $0.6 F_1$ shown.



Equation 1

$$P_r = P_t + G_t - L_{fs} - L_o - L_c + G_r$$

where P_r = received signal power in dBm

P_t = transmit power in dBm

G_t = transmit antenna gain in dBi

L_{fs} = free space loss in dB = $22 + 20 \log_{10}(d/\lambda)$, where d is the path length and λ is the radio carrier wavelength

L_o = all other path losses including diffraction loss and rain attenuation

L_c = cable and waveguide losses

G_r = receive antenna gain in dBi

path profile is the first step to ensuring the feasibility of the path. The profile indicates whether the path is free of obstructions. Because the atmosphere has a refractive index gradient that tends to bend radio waves back toward the earth, the radio horizon is usually greater than the optical horizon. The standard propagation model assumes an equivalent earth radius equal to $4/3$ the actual radius. To capture this effect on the path profile, a $4/3$ earth profile is used.

Before the advent of personal com-

puters, the designer plotted the profile on special graph paper. Today, the computer creates the $4/3$ earth profile when it plots the terrain profile, or equivalently, the terrain is plotted on a flat earth and the path is curved to show the effect of a $4/3$ earth. Sometimes anomalous atmospheric conditions occur and create the equivalent of a smaller earth radius. For this reason, some paths also are plotted assuming a worst-case $2/3$ earth radius.

Terrain usually is the dominant factor in the profile. Digital terrain databases are used to save the time and

trouble of extracting elevations from topographic maps. Generally speaking, a high-resolution database such as the 30-meter database produced by the U.S. Geological Survey is preferred.

Buildings and foliage also can obstruct the path, and these obstructions usually are not indicated on digital terrain databases. Therefore, a thorough site survey is required before the path is selected for construction. A typical path profile is shown in Figure 1.

Path loss prediction. A simplified link budget can be calculated using Equation 1, while the required signal amplitude for a particular path-availability can be found via Equation 2.

The lack of sufficient clearance above terrain obstacles will result in attenuation of the radio signal. Pure line-of-sight is not sufficient because all radio waves cast shadows, and a sufficient clearance above obstacles is required to ensure there is no diffraction loss. Clearance above terrain is measured in terms of the Fresnel zone radius, or F_n , where $n = 1, 2, 3$, etc. From diffraction theory, we can calculate the radius of the n^{th} Fresnel zone using Equation 3.

Usually, we are concerned only with the first Fresnel radius, F_1 . The path

Equation 2

$$P_{req} = -174 + 10 \log_{10}(B_n) + NF + C/N_{req} + FM$$

where B_n = equivalent noise bandwidth of the receiver in Hz

NF = receiver noise figure in dB

C/N_{req} = required carrier-to-noise ratio for a particular bit-error rate

FM = fade margin to account for fading and rain attenuation, usually 25 to 40 dB

Equation 3

where λ is the wavelength of the radio carrier

d_1 = the distance from the first end point to the point of interest

d_2 = the distance from the second end point to the point of interest

$$F_n = \sqrt{n \lambda \frac{d_1 d_2}{d_1 + d_2}}$$

✱ clearance must be at least $0.6 F_1$ to ensure negligible attenuation. At grazing ($0.0 F_1$), the attenuation can be as high as 15 dB, depending on the shape of the obstacle and its conductivity.

✱ The Fresnel radius reaches its maximum level at the center of the path, as shown in Figure 1. In the late 1950s, Ken Bullington published a nomogram that predicted the obstacle diffraction loss as a function of clearance. This nomogram is useful, but most designers simplify the problem to creating adequate clearance so that the predicted diffraction loss is zero, meaning that the clearance must be at least $0.6 F_1$.

A landline can be cut at an infinite number of points, but a microwave link can fail at only two spots.

Path availability. The path availability (also called link reliability) is the percentage of time that the received signal is above the required threshold,

P_{req}. It is sometimes expressed as the expected minutes of outage per year. The path availability is a function of the radio frequency, diversity (if any), fade margin, path length, and local climate. The International Telecommunication Union publishes reports with empirical models of required fade margin for different parts of the world. Designers use these reports to ensure adequate fade margins. A typical design requirement is 99.995%, which is equal to an expected outage of 26 minutes per year. ■

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✱ Ken Bullington, "Radio Propagation Fundamentals," Bell System Technical Journal, vol. 36, no. 3, 1957.

✱ International Telecommunications Union, TU-R P.530, "Propagation data and prediction methods required for the design of terrestrial line-of-sight systems," Feb 2007.

Jay Jacobsmeyer is president of Pericle Communications Co., a consulting engineering firm in Colorado Springs, Colo. He holds BS and MS degrees in electrical engineering from Virginia Tech and Cornell University, respectively, and has more than 25 years of experience as a radio frequency engineer.

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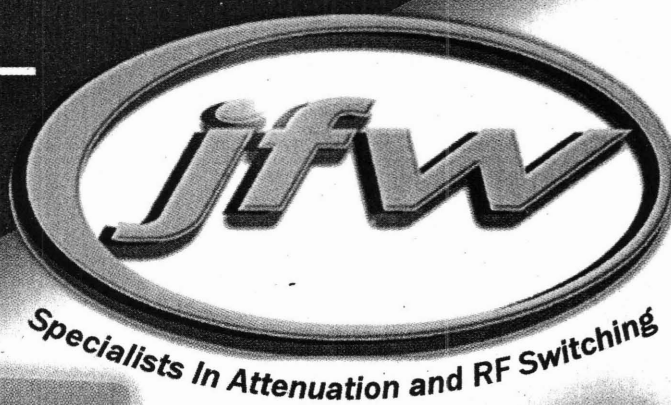


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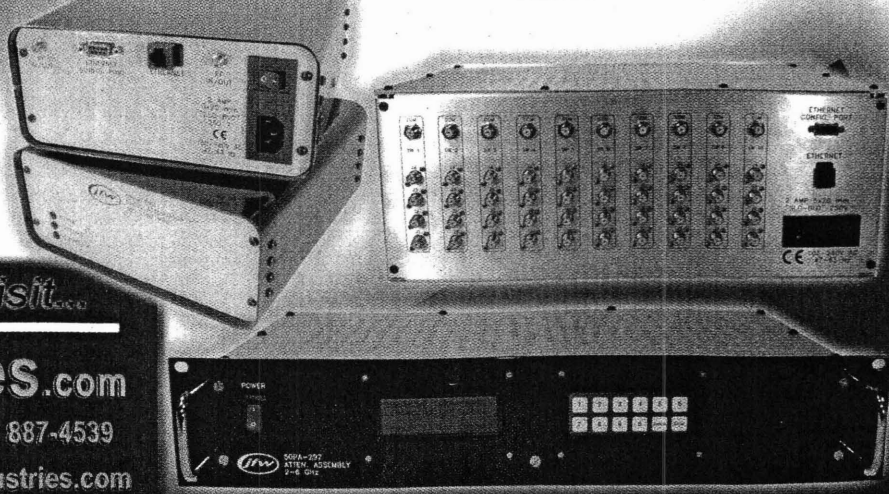
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Receiver front-end characteristics

By Harold Kinley, C.E.T.

The old slogan, "It's what's up front that counts," certainly holds true in the design of radio receivers. Under ideal conditions—low site noise, moderate-to-strong desired signal and no strong off-channel signals—almost any receiver will perform well. However, under less-than-ideal conditions, the design of the receiver front end becomes critically important. Design of the RF amplifier as well as front-end selectivity determines much of the operating capabilities of the receiver.

Wideband receivers

Wideband receivers designed for mobile use can cause significant problems when used as a base station receiver. If you are forced to use a wideband receiver for a base station, be aware of the limitations and weaknesses of such an arrangement. Many wideband mobile receivers have "wide-open" front ends. That is, there is little selectivity up front. Some wideband receivers use electronically tuned resonant circuits that use a varactor diode tuned by a control voltage as the receiver operating frequency is changed. Other receivers might provide one of more front-end "windows" that are changed or shifted for certain frequency ranges.

Intermod rejection

If you are comparing the specifications of a receiver with a wide front end with those of a receiver with a narrow front end, you might be a little misled. The front-end

selectivity is critical to good intermodulation performance. The more tuned stages ahead of the first active component (RF amplifier), the better the intermodulation rejection capability of the receiver will be, as long as the mixing signals fall out of the passband of the front end. If you look at just the receiver specifications, you might interpret

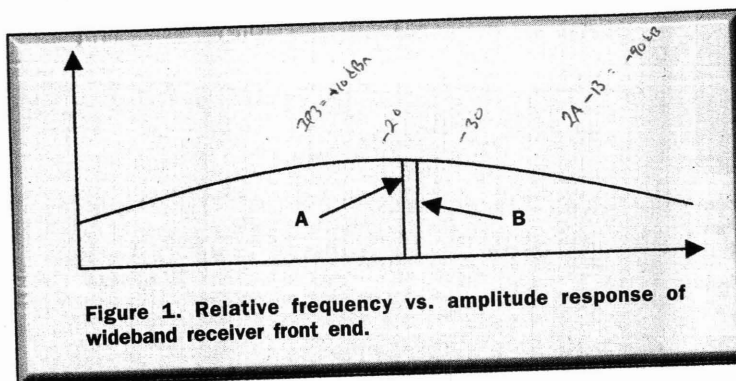


Figure 1. Relative frequency vs. amplitude response of wideband receiver front end.

that the intermodulation rejection figure of a specific wide front-end receiver is almost as good as another one with a relatively narrow front end. However, this doesn't tell the whole story. Suppose we are using a narrowband receiver and that the third-order intercept point for the RF amplifier is +10dBm. Further suppose that a signal at the first adjacent channel above the desired frequency is present at the amplifier input at a level of -20dBm. Let's call this signal A. (Refer to Figures 1-3.) Also present at the amplifier input is a signal (B) that is two channels up from the desired frequency at a level of -30dBm. Signals A and B will produce an intermod product ($2A - B$) at a level of -90dBm on the receiver frequency. When using the same signal inputs on a receiver with a wide front end, you might get similar results. However, a comparison of the receivers, using signal inputs far-removed from the center frequency, will show a significant difference in the results. Suppose for the narrow front-

will stay essentially the same, using the same signal levels at the input to the RF amplifier that were used in the previous example. However, the narrow front-end receiver will attenuate these out-of-band signals. Let's assume that signal A is down by 10dB to a level of -30dBm at the RF amplifier input and that signal B is down by 15dB to a level of -45dBm at the input to the RF amplifier. The level of the intermod signal at the output of the RF amplifier is down to -125dBm in the narrow front-end receiver. This intermod signal would cause much less trouble than the intermod signal of -90dBm in the wide front-end receiver. The bottom line is that receivers with wide-open front ends allow more signals to reach the input to the RF amplifier at a higher level. This increases the odds of intermodulation interference occurring.

Desensitization

Receiver desensitization occurs when a strong off-channel signal overloads a receiver front end and thus reduces the sensitivity to weaker on-channel signals. Receiver desense is usually measured by injecting an

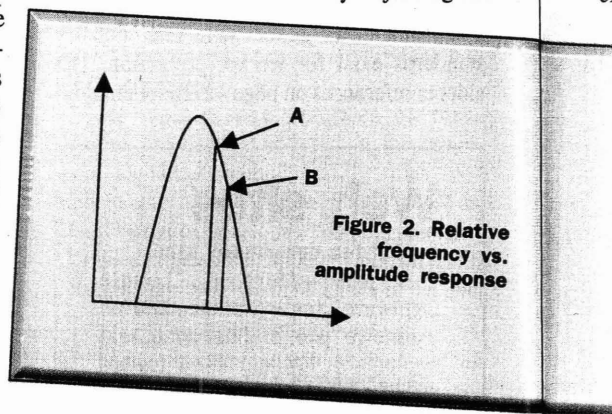


Figure 2. Relative frequency vs. amplitude response

on-channel signal to produce 12dB SINAD at the receiver output. This is the reference sensitivity. Then the signal generator is increased by 3dB to produce a better-than 12dB SINAD at the receiver output. Then an off-channel (one of the adjacent channels) signal at a

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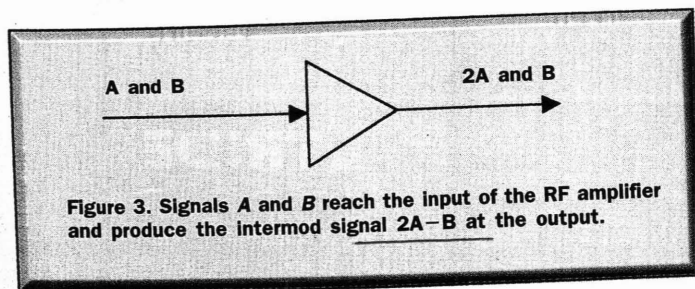


Figure 3. Signals A and B reach the input of the RF amplifier and produce the intermod signal $2A - B$ at the output.

end, receiver that the receiver is tuned to 152MHz and that signal A is 154MHz and signal B is 156MHz. Then the third-order intermod signal, $2A - B$, is equal to the receiver's tuned frequency, 152MHz. The wide front-end receiver

different modulating frequency is applied through a combining network to the receiver input. The off-channel signal is increased in amplitude until the SINAD meter again indicates 12dB SINAD. The difference in the level of this *undesired* signal and the *reference sensitivity* is the *desense figure* or *adjacent channel selectivity*. In one sense, this could also be called *dynamic range*, although this term is used to refer to a number of things regarding a receiver.

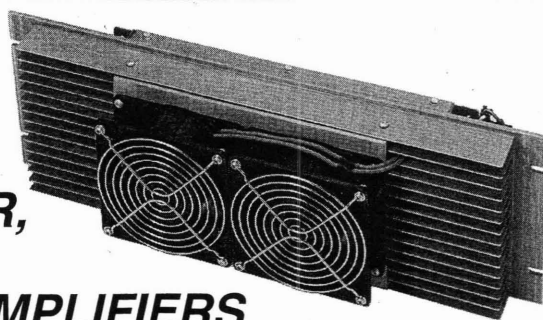
In a receiver with a wide front end, the undesired signal generator can be set several megahertz away from the on-channel (desired) signal generator and still cause desense. A practical example of this happened to one of the users of a repeater system. The repeater input was several megahertz down from the repeater output. While standing next to his mobile radio and using a walkie-talkie to "hit" the repeater, the user could not hear the repeater in the mobile radio until the walkie-talkie stopped transmitting. The walkie-talkie output was desensitizing the mobile receiver even though the receiver frequency was several megahertz away from the walkie-talkie output frequency.

Many times the phenomenon of receiver desense goes undetected. That is, the receiver never "hears" the desired signal and the operator might never be aware that there was a problem. In a mobile unit, the vehicle might have been near a powerful transmitter a few megahertz away from the receiver operating frequency, and the operator was never aware that desense was occurring in his receiver. When asked why he didn't respond to a call, he might answer that he didn't hear your call. This can create problems both from the personnel standpoint and technical standpoint. The technician is called to check his radio, and no problem is found. Is the receiver intermittent? Is the vehicle operator lying? Is the radio technician incompetent?

Parameters such as desense, intermodulation immunity and dynamic range all are interrelated.

The design of the receiver front end has a great bearing on all of these parameters. In receivers with a wide front-end design, users must be careful to reduce the possibility of overload from strong, off-channel signals. Avoid using a receiver with a wide front end as a base station—especially at a site with many local transmitters. If you must do so, then use a couple of bandpass cavities ahead of the receiver. If you are operating single-frequency simplex, it can help the site noise problem if you run the transmitter through the cavities also.

Until next time—*stay tuned!*



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Transmission lines: unravelling the mystery

By Harold Kinley

RF transmission lines have always presented an element of mystery. However, understanding the behavior of transmission lines is important to those of us who routinely work with these conduits of RF energy. Call them feed-lines, transmission lines or simply "coax"—the *purpose* is to get energy from a transmitter output to an antenna, or from an antenna to a receiver input.

Non-resonant lines

When a transmission line is terminated in its characteristic impedance, all

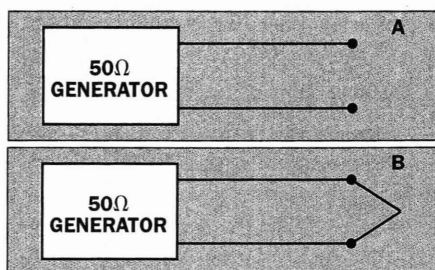


Figure 1. (A) A 50Ω transmission line that is unterminated, or open-circuited, at the end. The behavior of such a line is described in Table 1, below. (B) A 50Ω transmission line that is short-circuited at the end. The behavior of such a line is also described in Table 1.

of the power delivered to the load by the incident (forward-traveling) wave is absorbed by the load. The line is said to be *non-resonant* in this situation. The impedance at each point along the line, at any point between the generator and the load, is the same. There is no reflected wave on the line. The impedance will be equal to the characteristic impedance of the line. The length of the line in this case is not important, except where loss is concerned. You might say that, compared to resonant lines, non-resonant lines are boring.

Resonant lines

When a transmission line is terminated with an impedance that is *not* equal to the characteristic impedance of the line, things change. The impedance along the line between the generator and load varies but repeats at every half wavelength ($\lambda/2$) from the load. The load impedance may be purely resistive, or it may contain an element of reactance, either inductive or capacitive.

In Figure 1A at the left, a 50Ω transmission line (in a 50Ω radio system) is open-circuited at the load end. This is a theoretically perfect transmission line. It exhibits no loss and has a velocity factor of 1. Let's look down the line

from the load end toward the generator, or source, in *one-thirty-second-wave-length* ($1/32\lambda$) increments. We only need to cover $\lambda/2$ of line because the situation starts repeating at $\lambda/2$. The second column in Table 1 below shows what the impedance will be at each incremental point on the line. Remember, this is for an open-circuited line—a near-perfect open circuit.

Now, if we place a short circuit on the end of the transmission line shown in Figure 1B, the situation changes. The third column of Table 1 (for the short-circuited line) lists the impedance value at each location moving from the end of the line toward the generator. This is for a near-perfect short circuit.

Some interesting points can be gleaned by studying Table 1. First, for the open-circuited line, the impedance consists of infinite resistance and infinite reactance at the end of the line. As the distance increases down the line (toward the generator), the resistive component becomes 0Ω and the inductive reactive component decreases until it reaches 0Ω at $\lambda/4$ from the end of the line. Thus, $\lambda/4$ from the end of the line, the impedance has shifted from an open circuit to a short circuit—a complete impedance transformation.

Now, moving further down the line from the end, between $\lambda/4$ and $\lambda/2$, the resistive component remains at 0Ω while the reactive component switches from inductive to capacitive and increases up to infinity (∞) at the $\lambda/2$ point. It is interesting to note that the reactive component is equal to the characteristic impedance of the line at both the $\lambda/8$ and $3\lambda/8$ points. However, at the $\lambda/8$ point, the reactance is capacitive, but it is inductive at the $3\lambda/8$ point. At the $\lambda/2$ point, the impedance again becomes infinity ($\infty \pm j\infty$). Thus, the impedance repeats at $\lambda/2$. This entire sequence would repeat for every $\lambda/2$ section of the transmission line. So, it is only necessary to study $\lambda/2$ section of

Table 1. The first column represents the distance from the end of the line in $\lambda/32$ wavelength increments. As shown here, the impedance repeats at every half wavelength and inverts at every quarter wavelength. These figures represent the situation with a lossless transmission line.

DISTANCE FROM LOAD (λ)	OPEN-CIRCUIT LINE IMPEDANCE (Ω)	SHORT-CIRCUIT LINE IMPEDANCE (Ω)
0/32	$\infty \pm j\infty$	$0 \pm j0$
1/32	$0 - j250$	$0 + j10$
2/32 = 1/16	$0 - j120$	$0 + j21$
3/32	$0 - j75$	$0 + j33$
4/32 = 1/8	$0 - j50$	$0 + j50$
5/32	$0 - j33$	$0 + j75$
6/32 = 3/16	$0 - j21$	$0 + j120$
7/32	$0 - j10$	$0 + j250$
8/32 = 1/4	$0 \pm j0$	$\infty \pm j\infty$
9/32	$0 + j10$	$0 - j250$
10/32 = 5/16	$0 + j21$	$0 - j120$
11/32	$0 + j33$	$0 - j75$
12/32 = 3/8	$0 + j50$	$0 - j50$
13/32	$0 + j75$	$0 - j33$
14/32 = 7/16	$0 + j120$	$0 - j21$
15/32	$0 + j250$	$0 - j10$
16/32 = 1/2	$\infty \pm j\infty$	$0 - j0$

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Kinley's email address is hkinley@home.com.

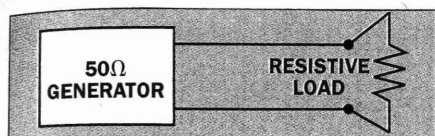


Figure 2: A 50Ω transmission line is terminated in a purely resistive load that is unequal to the characteristic impedance of the transmission line. See text for examples.

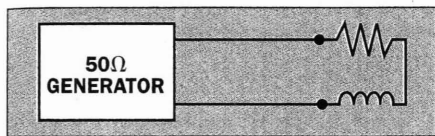


Figure 3: This 50Ω transmission line is terminated in a complex impedance load consisting of a 25Ω resistor and an inductive reactance of 25Ω. See text for details.

line to understand what is happening on the entire line.

The third column of Table 1 summarizes what happens with a short-circuited transmission line. Of course, at the short-circuited end, the impedance is $0 \pm j0\Omega$. Between the end of the line and the $\lambda/4$ point, the reactance is inductive. This is the reverse of the open-circuited line. Again, at the $\lambda/8$ and $3\lambda/8$ points, the reactance is equal to the characteristic impedance of the line. At the $\lambda/4$ point, the impedance becomes an open circuit ($\infty \pm j\infty$). Between the $\lambda/4$ and $\lambda/2$ points, the reactance switches to *capacitive* reactance. And, at the $\lambda/2$ point, the short circuit at the end of the line is repeated and will do so at every multiple of $\lambda/2$ down the line toward the generator.

From this it can be concluded that for an open- or short-circuited transmission line, the impedance at any point between the end of the line and the $\lambda/2$ point will consist of 0Ω resistance and a reactive component that extends from 0 to infinity. Note that for an open-circuited line, the resistance component switches abruptly from 0Ω to infinity at the $\lambda/2$ point and every multiple thereof referenced to the end of the line. Conversely, for a short-circuited line, the resistance component switches abruptly from 0Ω to infinity at the $\lambda/4$ point and every odd multiple thereof referenced to the end of the line.

It can further be stated that if the load impedance is *purely resistive* and *greater than* the characteristic impedance of the line, the reactance will be *capacitive* at any point between the end of the line and $\lambda/4$ from the end, and *inductive* between the $\lambda/4$ and $\lambda/2$ points. If the load impedance is *purely resistive* and *less than* the characteristic impedance of the line, the reactance will be *inductive* at any point between the

end of the line and the $\lambda/4$ point, and *capacitive* between the $\lambda/4$ and $\lambda/2$ points.

Figure 2 at the left illustrates a transmission line, with a characteristic impedance of 50Ω, terminated with a load that is a pure resistance of 25Ω ($25 \pm j0$). The VSWR will be 2:1. We can determine the impedance at $\lambda/4$ from the end of the line by using the following formula:

$$Z_1 = \frac{Z^2}{Z_L}$$

where Z is the characteristic impedance of the line, Z_1 is the impedance at the $\lambda/4$ point and Z_L is the load impedance. In Figure 2, the load impedance is a pure resistance of 25Ω and the characteristic impedance of the line is 50Ω. Now, plugging these values into the equation above we get:

$$Z_1 = \frac{50^2}{25} = \frac{2,500}{25} = 100$$

Thus, the impedance at the $\lambda/4$ point is 100Ω. The VSWR will still be 2:1. Assuming a lossless line, the VSWR at any point on a line will be the same.

If in Figure 2 the 50Ω transmission line was terminated by a load with a pure resistance of 100Ω, the VSWR would still be 2:1. The impedance at a point $\lambda/4$ away from the end (according to the formula above) will be:

$$\frac{50^2}{100} = \frac{2,500}{100} = 25$$

Suppose the line is terminated in a complex impedance, one that contains both resistive and reactive components. The formula still holds true, but now we must enter the load impedance in complex form. In Figure 3, a 50Ω transmission line is terminated in an impedance of $25 + j25$ ohms. Plugging this into the formula we get:

$$Z = \frac{50^2}{(25 + j25)} = \frac{2,500}{(25 + j25)} = \frac{2,500}{25(1 + j)} = \frac{100}{(1 + j)}$$

$$= \frac{100}{(1 + j)} \times \frac{(1 - j)}{(1 - j)} = \frac{100(1 - j)}{2} = 50 - j50$$

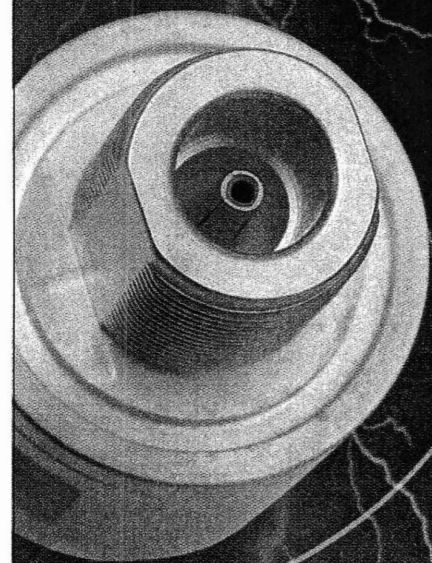
Thus, the impedance $\lambda/4$ down the line is equal to $50 - j50\Omega$. The inductive reactance present in the load impedance has been transformed into capacitive reactance at the $\lambda/4$ point down the line. To reiterate, *although the impedance on the line changes, the VSWR does not change.*

Next month's column deals with practical applications of this information.

So, until next time ... *stay tuned!* ■

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How's your antenna?

Familiarity with antenna characteristics helps predict performance in matching the right 'stick' to the right application.

By Patrick E. Buller

The difference in the light observed at the center of the parabolic beam and that of the bare bulb is *gain*

To the casual observer, it's just a wire or just some sort of metallic nuisance that needs painting. We recognize it as an antenna, a metallic device, such as a rod or wire, for radiating or receiving radio waves. An antenna is most efficient when its physical length is appropriate for the frequency to be operated. The physical length is determined by its relation to *wavelength* (λ), which in turn is related to the speed of light.

The velocity of light, V , is roughly 186,000mi./s. Multiplying V by 5,280ft./mi. yields 984×10^3 ft./s. *Megahertz* is a million events per second. One wavelength is also one hertz, which is also equal to one period of events per second. Therefore, 984,000,000 divided by frequency equals one wavelength (1λ).

One wavelength = $984 \div F(\text{MHz})$. For the $\frac{1}{2}\lambda$ dimension, in feet, 492 is the value to be divided by the frequency, and $492 \div F(\text{MHz}) = \frac{1}{2}\lambda$, in feet.

Example: $984 \div 155\text{MHz} = 6.35\text{ft.} = 1\lambda$.

$\frac{1}{2}\lambda = 3.175\text{ft.}$

$3.175\text{ft.} \times 12 \text{ in./ft.} = 38.1\text{in.}$

$\frac{1}{4}\lambda = 19.05\text{in.}$

All of the above is relative to the speed of light. If we are going to use wires or conductors of any type as the antenna, then things take on a different characteristic. Anything other than free space will slow down energy, and the ratio of free-space travel to conductor travel is the velocity factor, usually noted as vp.

The *isotropic* antenna is a fictitious antenna, also called the *point source*. It is analogous to a bare flashlight bulb theoretically illuminating a room evenly. When the flashlight bulb is placed at the focal point of a parabolic reflector, a light beam is formed. The difference in the light observed at the center of the parabolic beam and that of the bare bulb is *gain*. This comparison relates directly to microwave antennas that use a parabolic dish. In fact, microwave antennas are compared to the isotropic antenna, and all gains are written as units of *dBi*, meaning decibels relative to a hypothetical isotropic antenna.

By definition, the *dipole* antenna is a $\frac{1}{2}\lambda$ conductor, with energy supplied in the center, as shown in Figure 1 on page 49. It is a balanced-feed device, and its parameters are measured or compared at a height above ground of $\frac{1}{4}\lambda$. Its gain, compared to the isotropic antenna, is 2.15dBi. This antenna is easily built and calibrated, and it is the standard antenna for measurements of frequencies below 900MHz. Antennas with non-dBi gain figures are given in units of *dBd* (i.e., referenced to the dipole). Figure 2 on page 49 shows the voltage and current relationship of a $\frac{1}{2}\lambda$ dipole antenna.

Let's look at an antenna for its feed point impedance over several wavelengths. Keeping the physical length constant, the frequency will be changed and the base impedance measurements can be taken and the results plotted as shown in Figure 3 on page 49. A horizontal line that represents 0Ω resistance at the left

and approaches infinity at the right represents the resistance. Inductive reactance lies above the horizontal line, and capacitive reactance lies below.

In the electronics world, we like to have everything resistive. So, looking at the spiral, we determine that if we operate at the point where the spiral line crosses the resistive line, the antenna would then be resonant. (The energy boys call this point unity power factor.)

So, is the free-space formula in error? Not really, it just needs some modification. In reality, the frequency that provided this zero reactive point was plugged into the formula to find a "K" factor that would modify the free-space formula, $492 \div F(\text{MHz})$, to bring us to the zero point. That "K" factor is 0.95 for wire antennas up to about 30MHz. Beyond 30MHz, the diameter of the wire also affects the overall length for a pure resistive impedance. The result is the classic $468 \div F(\text{MHz})$ seen in most text books for a $\frac{1}{2}\lambda$ antenna.

Figure 4 on page 50 shows the impedance spiral of a $\frac{1}{2}\lambda$ antenna cut to the dimensions of $468 \div F(\text{MHz})$. Again, this is only good for wire antennas or thin elements for the lower frequencies and at $\frac{1}{4}\lambda$ above ground. The lower the antenna is to ground, the lower this impedance will become. A rule of

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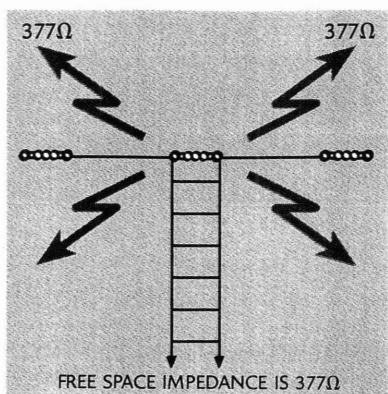


Figure 1. An antenna is a special case of a terminated transmission line.

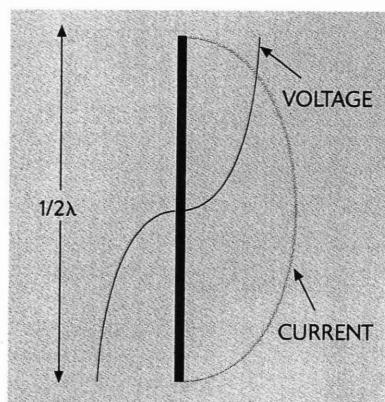


Figure 2. Voltage and current relationship of a $1/2\lambda$ antenna.

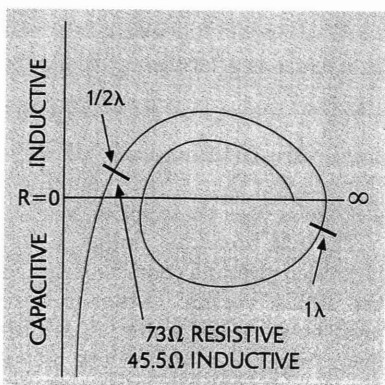


Figure 3. Impedance spiral of $1/2\lambda$ antenna. Free space dimension $492 \div F(\text{MHz})$.

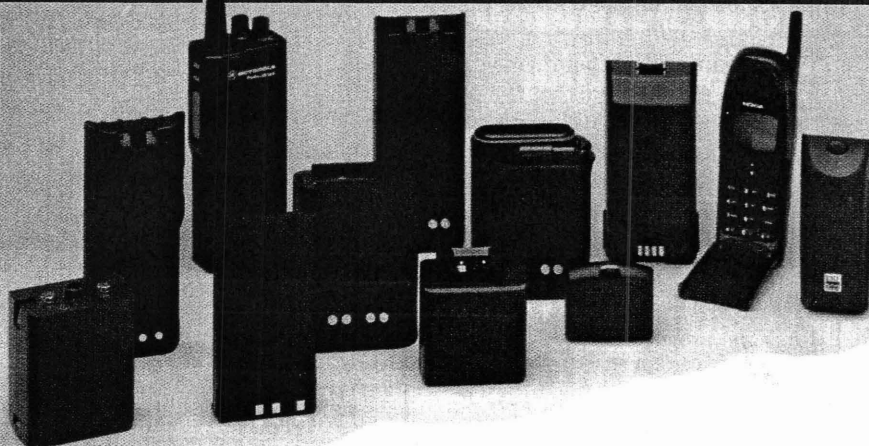
thumb is if the antenna is about $1/8\lambda$ above ground, the impedance will measure somewhere around 50Ω. In the amateur radio band, this means an 80m antenna up 30' will closely match the magic number of 50Ω. Also note that the antenna has a higher impedance in the center if it is 1λ long. At $3/4\lambda$, the center impedance is again low.

That means a center-fed antenna can and will operate on the 3rd harmonic. Examples would be a $1/2\lambda$ on 7MHz and a $3/4\lambda$ on 21MHz.

Figure 5 on page 50 shows the impedance spiral of a $1/4\lambda$ antenna cut for free-space dimensions. Note that the numbers are exactly one-half of the $1/2\lambda$ dipole: $Z_0 = 36 \times$

$j21\Omega$. Again, the element must be shortened to yield pure resistance, using the same formula except $1/2$ of the total length. The only difference between this antenna and the dipole is that the ground plane is considered infinite in length. In reality, it need only be $1/2\lambda$ diameter in all directions. The $1/4\lambda$ antenna is a low-

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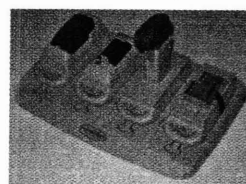
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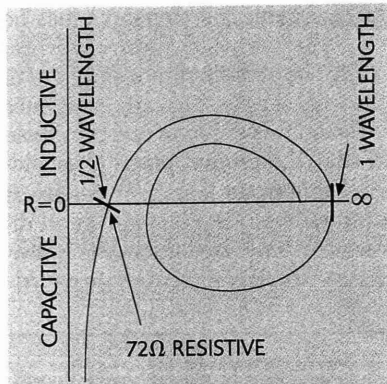


Figure 4. Impedance spiral of a $\frac{1}{2}\lambda$ antenna shortened by $K=0.95$ or length= $468 \div F(\text{MHz})$.

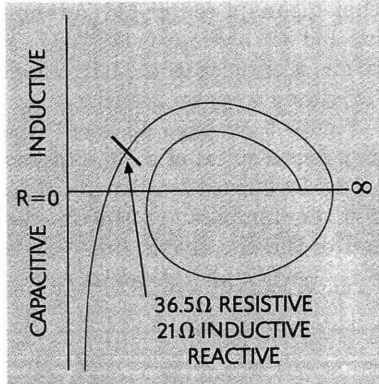


Figure 5. Impedance spiral of a $\frac{1}{4}\lambda$ antenna free space dimension $492 \div F(\text{MHz})$.

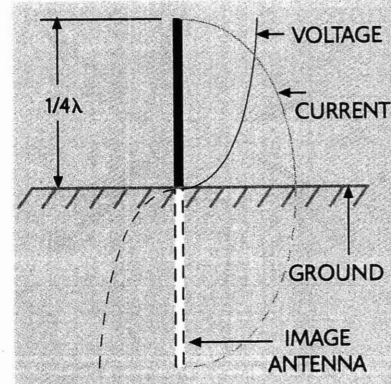


Figure 6. Voltage and current relationship of a $\frac{1}{4}\lambda$ antenna.

impedance, or current-fed, antenna, as shown in Figure 6 above. It has a voltage minimum at the feed point and again at the $\frac{3}{4}$ point. Advantages of a $\frac{3}{4}\lambda$ antenna are that the impedance can be near 60Ω (a better match than the $\frac{1}{4}\lambda$), and a minor lobe

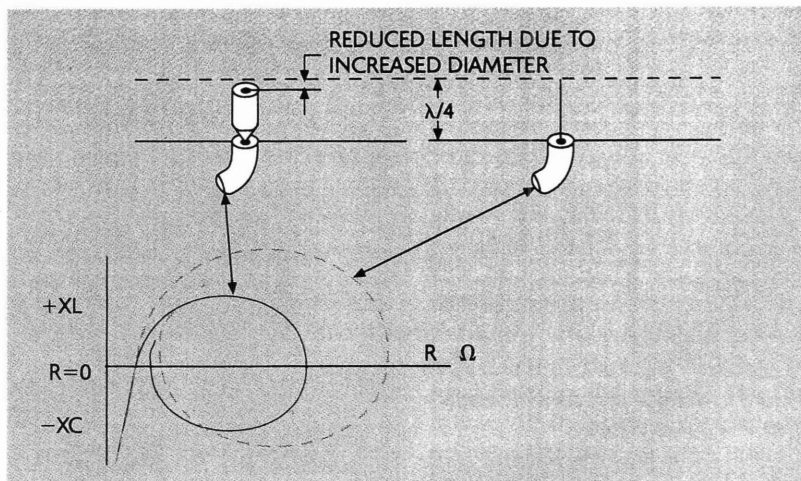


Figure 7. Impedance spiral of a $\frac{1}{4}\lambda$ vertical antenna as a function of conductor diameter often referred to as bandwidth.

appears above the horizon, which is helpful with radiation to higher sites.

Figure 7 shown above amplifies another consideration of the radiating element. Large diameters lower the antenna's quality factor, or "Q." In other words, it makes it a broadband device. There is a larger spread of frequencies with which the antenna will operate with acceptable standing wave ratio. This is primarily due to the amount of capacity added because of the increased area. To keep the antenna resonant, the conducting element must be shortened to compensate for the added capacity. This brings the topic of the ratio of diameter of radiating element to wavelength dimensions, as shown in Figure 8 on page 51. In other words, the "K" factor is the determinate when the antenna element is larger than a

small wire. In the case of VHF, anything larger than a $\frac{1}{8}$ rod must use a K factor less than 0.95.

The AM broadcast industry first identified antenna characteristics in the 1920s. Vertical antennas were used because the radiation is uniform from the center of the tower. Dipole antennas have directivity, with nulls at the ends of the conductor. The broadcasters' interest was to obtain the most signal strength, so the vertical antenna underwent a series of tests to determine the optimum height. Figure 9 on page 51 graphs signal strength measured with different vertical antenna heights and 1kW of power. Note that maximum power is received with the antenna at 0.625λ , and almost nothing at one wavelength.

Figure 10 above graphs a side view of radiation with different

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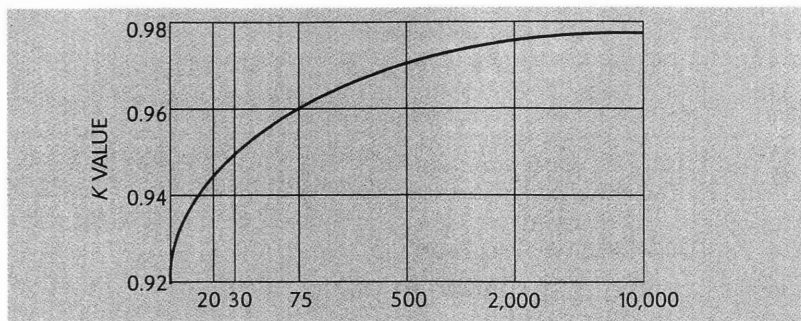


Figure 8. Ratio of halfwavelength to conductor diameter—same units.

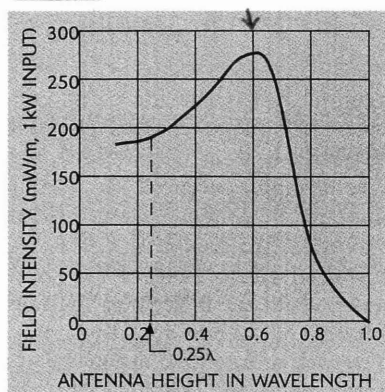


Figure 9. Plot of field strength vs. vertical antenna length.

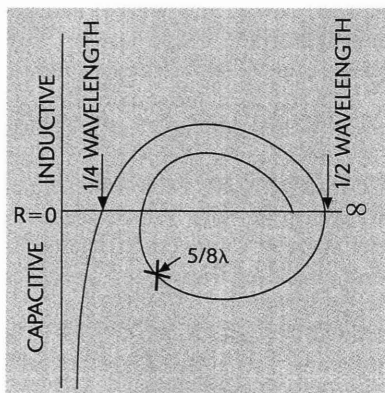


Figure 11. Impedance spiral of a vertical antenna. Note low resistance and capacitive reactance. Series inductance cancels. This is how the Larsen mobile $5/8\lambda$, 3dB gain antenna came about.

antenna heights. At the $5/8\lambda$, you can see there is a minor lobe beginning at the 75° point, that gets somewhat larger at $3/4\lambda$ (not shown). Installing a $3/4\lambda$ antenna is an attractive prospect for vehicles because this antenna not only radiates in the horizontal but also has an elevated signal that can bounce off buildings. The $3/4\lambda$ is particularly effective for motorcycle

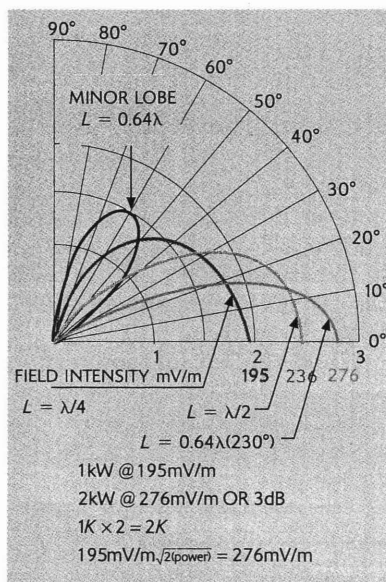


Figure 10. Vertical plane patterns for three values of antenna height. Field intensity is millivolts per meter.

radios using the VHF band. The impedance of a $3/4\lambda$ vertical is closer to 50Ω than the $1/4\lambda$ antenna.

The late James L. Larsen, founder of Larsen Antennas, introduced the $5/8\lambda$ antenna for mobile use by incorporating a coil at the feed point that cancels the capacitive reactance, leaving the antenna just above 50Ω resistive. It is marketed as the 3dB gain VHF antenna.

Another example of antenna impedance vs. height is shown in Figure 11 above. Any multiple of $1/2\lambda$ results in high impedance, sometimes referred to as voltage feed. Multiples of $1/4\lambda$ are referred to as current feed.

It should be pointed out that even a small piece of wire is a high-impedance device, if it is measured with an ohmmeter. It is only low resistance when it's employed as a resonant device.

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How's your antenna?

Part II—The halfwave antenna is expanded into many other configurations, each having specific contributions to how RF energy is sent or received.

By Patrick E. Buller

In the past, antennas were generally used for fixed, single-frequency operations. Today's needs require antennas to have a much broader bandwidth to cover more frequencies. These new requirements include repeaters with wide frequency separation or the combining of more than two transmitters into one antenna.

One approach to increased bandwidth is the incorporation of larger-

diameter material that lowers the antenna "Q" (discussed in Part I). Methods used to accomplish this incorporation include increasing conductor diameter or using more than one paralleled conductor. Parallel conductors define a circle that has the effect of a single conductor with a diameter equal to the spacing of the parallel conductors.

The bow tie antenna is another example of a large surface area that also increases bandwidth, often used in antennas for 450MHz and above. The folded conductor of a bow tie antenna also changes impedance.

Impedance change

In a classic, folded dipole antenna, as shown in Figure 1 at the left, there are parallel conductors, but only one conductor is fed power. It raises the impedance of the feed point by the relationship of quantity of conductors, squared, times the antenna impedance. For example, a pair of wires $(2)^2 = 4 \times 72\Omega = 288\Omega$.

A three-conductor antenna raises the impedance nine times, providing all conductors are the same diameter. If diameters are equal, then spacing between wires and conductors is not a factor. If the diameters of the conductors are not the same, the spacing is indeed important.

If, instead of feeding only one of the two conductors, both conductors of each leg are made common, as shown in Figure 2 at the left, then the bandwidth is the same as the folded dipole, but the impedance is close to the classic 72Ω . Why the impedance difference? The folded dipole shares the current equally between the parallel wires; therefore the current in each wire is half that of a single conductor.

The folded dipole has several features that make it useful as a vertical collinear antenna. Besides increased bandwidth, wind loading is somewhat increased, but vibration is drastically reduced, which eliminates metal fatigue that is common in single elements of ground-plane antennas. Noise caused by particles whizzing past during high winds is reduced, and the lack of sharp points eliminates corona noise. Corona noise is observed by hearing a high-pitched squeal sounding like a siren with ever-changing pitch. Corona noise stops immediately after a lightning strike.

Impedance changes when different diameter conductors are part of the feed. The final impedance is a function of the ratio of diameters of each element and the spacing between conductors.

When comparing the impedance of a center-fed dipole of 72Ω to that of a ground-plane antenna of 36Ω , consider the geometry. The dipole is two conductors in the same plane, and the vertical antenna has one vertical, the other horizontal. Moving the horizontal element from horizontal down to about 40° raises the impedance to around 50Ω , which is valid for a single-element vertical antenna with a modified ground plane as shown in Figure 3 at the left.

The collinear antenna

Collinear means having axes

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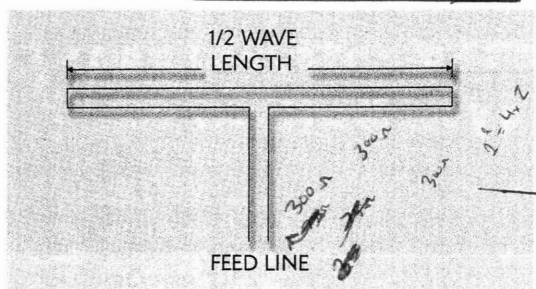


Figure 1. One halfwave antenna.

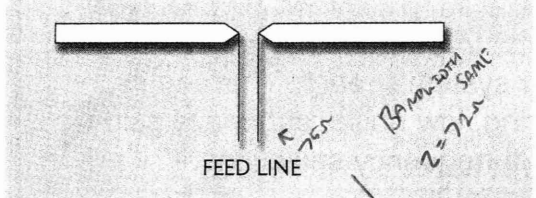


Figure 2. One halfwave antenna.

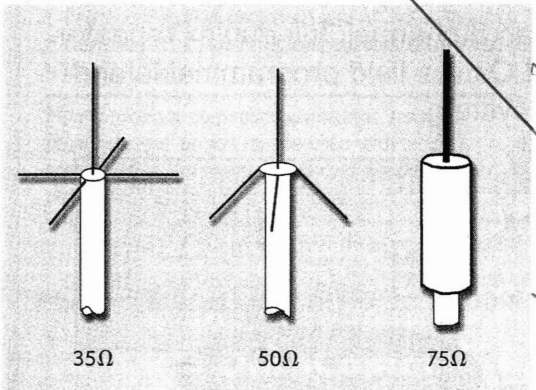


Figure 3. Impedance change by angle.

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lying end to end in a straight line. Collinear antennas for the two-way radio market are usually made up in pairs. This is a gain of 3dB when one more element is added to a single element. For an additional 3dB, the two elements must be doubled, yielding four elements for a total antenna gain of 6dB.

The most common 6dBd vertical antenna for the 150MHz band consists of four halfwave rigid coax cable pieces fed at the bottom with an impedance-matching transformer enclosed in a fiberglass tube as shown in Figure 4 at the right. Note that the coax antenna elements have conductors changing positions between elements. All elements must have the same phase relationship for each element to contribute energy perpendicular to the elements. Therefore, it is necessary to have correct polarity. Remember that the phase changes 180° in halfwavelength. If the polarity were not changed, and elements were not transposed, the antenna would be two wavelengths long. No energy would radiate perpendicular to the elements, meaning no signal on the horizontal—just a four-leaf clover pattern with no radiation broadside or off the ends. In reality, the coax linear is not a true 6dBd but about 5.4dBd because of the uneven power distribution. The bottom element has to feed the other three, and the top element feeds none.

Another common antenna in the two-way industry is the folded dipole array. It is fed with equal lengths of coax cable, which allows energy to arrive at all antenna elements at the same time. Figure 5 at the right shows a cable arrangement that is not only acting as a phasing harness but is impedance matching as well. Quarterwave transmission lines will also perform as transformers. The advantage of this type of antenna, compared with the coax collinear, is each element is fed the same amount of power and its gain is near the full 6dBd. Arranging the elements around the vertical support can modify the radiation

pattern. This option is not available with the coax collinear antenna.

Downtilt

Collinear VHF antennas with a 12° beamwidth that are installed in high elevations do not cover nearby low elevations. In short, these antennas talk well mountaintop-to-mountaintop, but not down to the highway, where they are needed. Downtilt is when the beam is electrically lowered below the horizontal plane. This is done by a phase shift of power between each radiating element. This phase difference between elements forces the beam down from the horizontal.

Figure 6 on page 38 shows the phasing required for downtilt. Each element going in the vertical direction has leading phase from the lower element. The downtilt is limited to about 6°. Beyond this limit, side lobes appear that defeat the goal of downtilt and gain.

Phased arrays

A single antenna has a definite gain. Duplicating this antenna and feed from the same source will provide 3dB more gain. Duplicating again will add 3dB more. The collinear antenna is one example of phased array. In the classic broadside phased array as shown in Figure 7 on page 38, eight dipole antennas are connected to a transmission line. Note that the transmission line is transposed. To have the antenna appear as one large antenna, all elements need to be fed with the exact same phase. For example, assume all elements will have the left side instantaneously positive. Each antenna is spaced halfwave apart, therefore the transmission line is transposed because the energy is shifted 180° in the halfwave transmission line. The center transmission line, however, is not transposed because the energy being fed in the center arrives at each antenna at the same time. If the antenna array were fed from the bottom, the center phasing harness would have a transposition.

The key in any phased array is

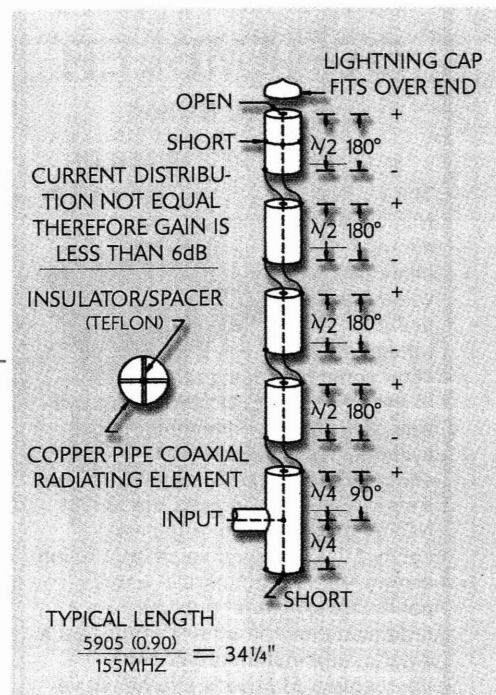


Figure 4. End fire collinear antenna.

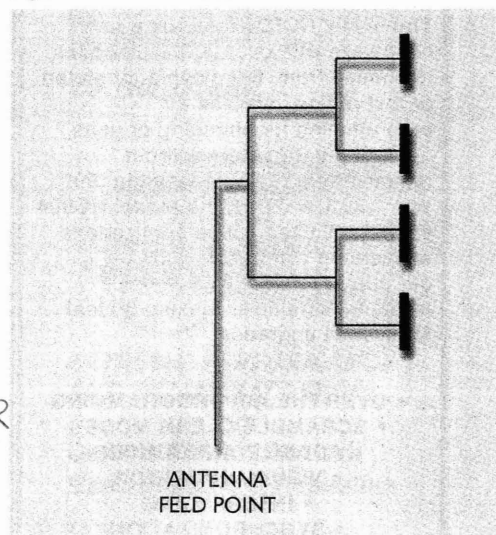


Figure 5. The collinear.

keeping all elements in the correct phase relationship. Antenna patterns can also be shaped by changing the spacing of antenna elements and different feed phase angles.

Log periodic antennas

A log-periodic antenna, (as shown in Figure 8 on page 39) is a variation of the broadside antenna. The radiating elements decrease in length according to the logarithmic function. The design is based on

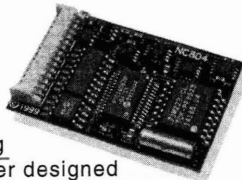
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antenna characteristics that are periodic with the logarithm of the frequency. The variation of the characteristics is small over a single time period, allowing the antenna to operate frequency-independent. In other words, it has an acceptable VSWR over a 2:1 frequency change—the detracting feature is that it has less gain than a Yagi-Uda antenna having the same number of elements, but it is less susceptible to ice buildup. Note the phase reversal of the feed line at each element.

Parasitic array

The multi-element Yagi-Uda antenna was described by S. Uda in Japanese and subsequently in English by H. Yagi. It became first known in the United States as the yagi antenna. A three-element antenna is shown in Figure 9 on page 40 with optimum spacing for forward gain. The longer element is the reflector, and the driven element (that takes power) and the director are shorter. The driven element is the base where all elements are derived from. Its length is found by $492(K/F(\text{MHz}))$, where K is the ratio of diameter to wavelength. The reflector is 5% longer and the director is 5% shorter than the driven element. Optimum spacing is 0.2 wavelengths from reflector to driven element and 0.1 to 0.15 wavelengths between the driven element and the director. A single reflector is all that is required, but additional directors can be added for more gain. Some antennas may have as many as 48 directors, but the contribution of any more than a dozen is so small, it is best to add another full antenna and feed it in phase. The exception would be when insufficient real estate exists to add antennas side by side.

The operation of a Yagi-Uda antenna is based on all elements accepting energy from the driven element and re-radiating in the correct time relationship. Consider a single wave radiating from the driven element. The reflector receives this energy, and it begins to radiate. It does so in the correct time relation-

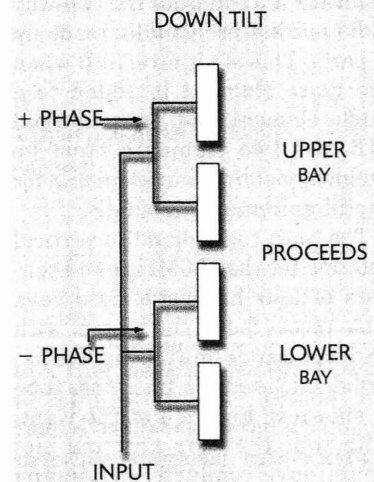


Figure 6. Downtilt.

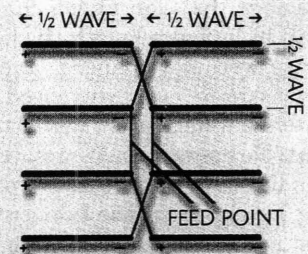


Figure 7. Broadside array.

ship to be in phase with the next wave produced by the driven element. The same contribution is experienced for the directors. They all add in-phase in the forward direction but add out-of-phase in the reverse direction. Anything that modifies the phase relationship directly affects the forward gain and its front-to-back ratio. Front-to-back ratio is the energy sent in the forward direction compared to that measured at 180°, and it is expressed in decibels. Ice buildup is one of many factors that alters the performance of the antenna. To make a Yagi-Uda antenna broadband, use large diameter elements, and no more than five. The more directors used, the higher the

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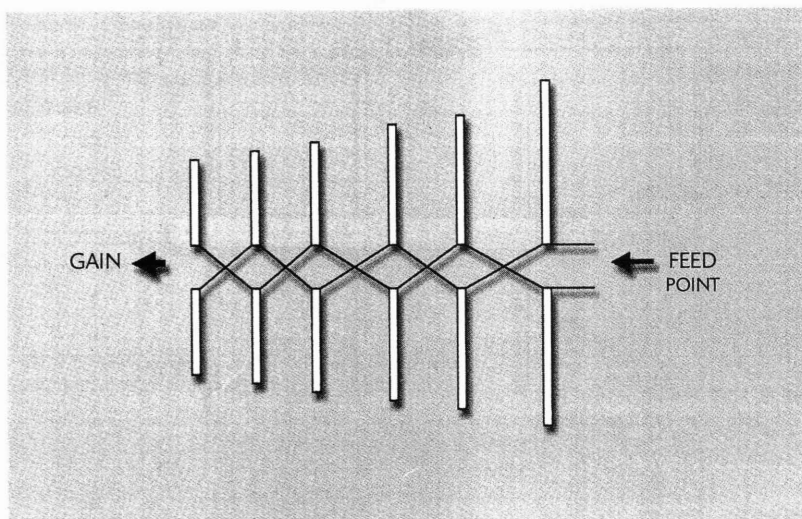


Figure 8. Log periodic antenna.

✱ antenna Q is, resulting in less bandwidth.

Slot antennas

A slot antenna is the conjugate of a wire antenna. Figure 10 on page 40 shows a large conductive sheet where a halfwavelength section of material is cut out. The material removed is a half-dipole antenna. The slot, then, is a halfwave antenna, but with reverse properties. Its high impedance point is in the center compared to the dipole with the lowest impedance at the center. Polarization is vertical, and it is a balanced antenna.

✱ When the antenna is rotated to where the slot is vertical, it is now radiating horizontal. Gain is achieved by stacking several in the vertical plane, one above the other. This antenna is the workhorse for the TV broadcast band. It is immune to lightning because its feed point is inside a grounded structure. Another use for the slot antenna is with guided missiles, aircraft and counterintelligence operations where the metal skin of the vehicle is used as the antenna. The slot is filled in with non-conductive resin, painted and completely camouflaged. If the slot antenna is backed with a cavity, it becomes highly directional and does not radiate inside the vehicle. Some texts incorrectly refer to this product as a "pylon antenna."

Parabolic antennas

Figure 11 on page 40 shows a parabolic antenna. It behaves identical to the flashlight. A source of energy at the focal point is re-radiated in the forward direction. The focal point can consist of a simple two-element Yagi-Uda antenna, a horn or an inverse parabolic antenna fed from the center of the larger antenna. The latter is common in space research antennas. For point-to-point microwave, the first two "illuminators" are common. Gain is calculated as $G = 20\log D + 20\log F - 52.5$, where D is diameter in feet and F is in megahertz. The industry classifies this antenna at 55% efficiency. If a radome is installed, then the gain figure must be reduced to account for the radome loss.

If high-performance antennas are required, a shroud with RF absorbing material is added around the dish to attenuate most of the side-lobe energy and to improve the front-to-back ratio. To the observer, the feed horn is further shielded except for the front of the dish. The reflector can be either solid, grid or mesh screen, depending on the frequency used. The higher the frequency, the tighter the tolerance must be in dish construction. As long as the opening in the material is less than 0.01 wavelengths, it appears as a solid. Vandalism can be a problem because parabolics present attractive



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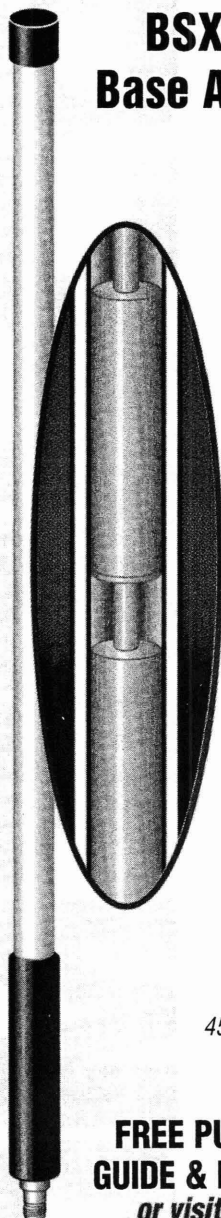
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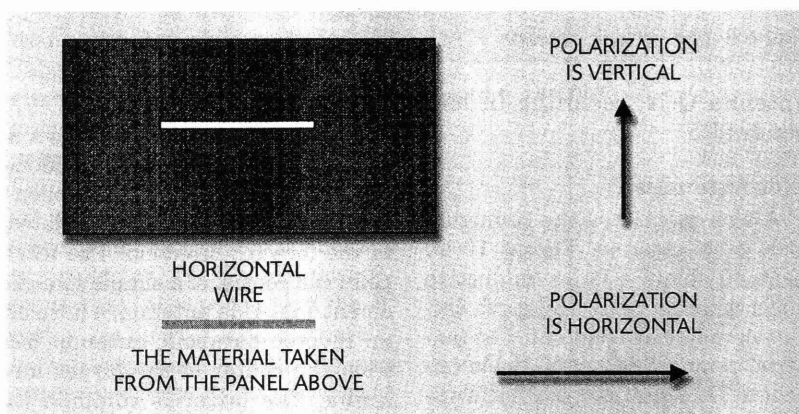
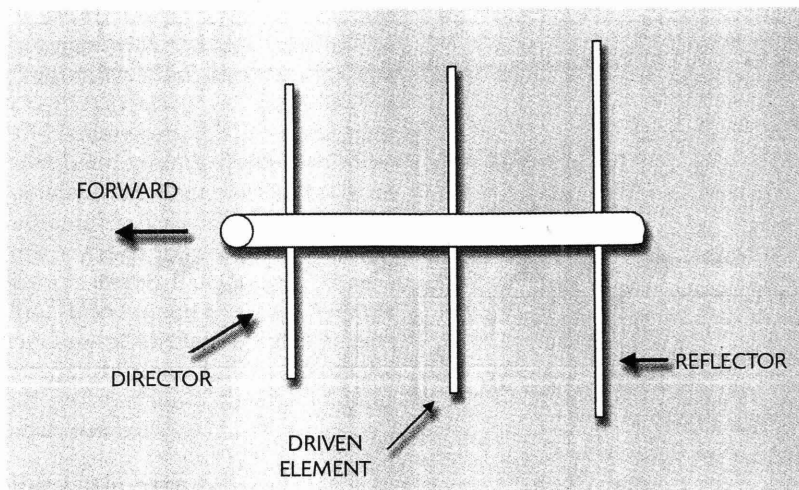
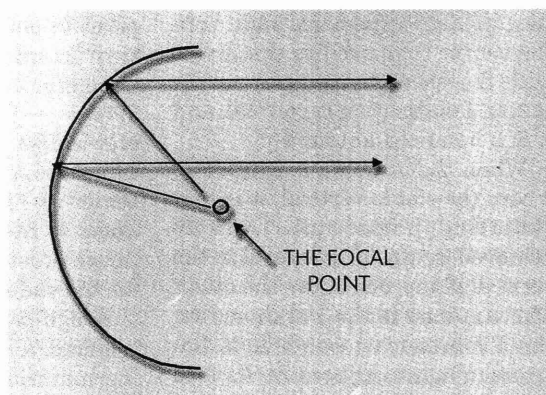


Figure 9 (top). Yagi-Uda antenna. Figure 10 (middle). The slot antenna has polarity opposite that of the wire. Figure 11 (right). The focal point is where all energy is reflected to. It is a narrow area and is easily missed by damage, often caused by leverage on the rear of the antenna such as waveguide.



targets. At 6GHz, bullet holes contribute little distortion. It's when the feed horn or waveguide is hit that the problem becomes serious. An example is a shipboard radar antenna. Most are constructed of metal screen to reduce wind loading with little reduction in either gain or beam width. Along that line, most navigational radar antennas have little material vertical but large in the horizontal. The large horizontal focuses the beam to be narrow

in the horizontal and large in the vertical. This not only allows accurate bearing measurements but also allows the vessel to rock about vertically and not lose its target. A pencil beam is required for pinpoint accuracy such as fire control radar or speed measuring devices. ■

Part I of "How's Your Antenna" appeared in the August 2001 issue of *Mobile Radio Technology*.

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The complex world of wireless E911

Ensuring that various technologies work seamlessly is critical when lives are at stake

Locating callers in an emergency is much more difficult in wireless than in the traditional wired world. After all, each wired phone line connects to a physical location, so finding the address is simply a matter of searching a database using the phone

anywhere in the coverage area.

However, locating the caller is only the beginning. The position information must then be routed to the correct public-safety entity to dispatch rescuers, passing through a number of hands to get there. In this article, we ex-

(PSAP). Ensuring these technologies work seamlessly is critical during an emergency when lives are at stake.

Locating wireless 911 callers is governed by the FCC's enhanced 911 mandate. According to the FCC's regulatory framework, network-based and handset-based location technologies may be used. Accuracy requirements differ for the two. We'll focus on a Polaris-developed network-based technology, wireless location signatures (WLS), which does not require changes to the handsets or base stations. Using this new software-only technology to locate callers enables the service provider to rapidly and economically roll out E911 service because there is no need to wait for new handsets to penetrate the market or new radio hardware to be installed in the network.

Regarding the transportation of location information, we'll focus on the world's most widely used air interface, GSM, to illustrate the critical flow of information during an emergency call. In GSM, there are mobile location centers (MLCs) to manage the process of locating the caller and notifying the PSAP. These MLCs must interact with both the radio network and equipment at the PSAP to coordinate the request and deliver the caller's location.

The E911 location service uses a layered model (Figure 1). The foundation level is the position determination layer. This is where the actual estimation of the caller's location takes place. In many ways it is analogous to the physical layer in an air interface. This is where the actual location technology comes into play: WLS, assisted global positioning system (AGPS) or time difference of arrival (TDOA). The position determination layer then hands the location information over to the

Figure 1

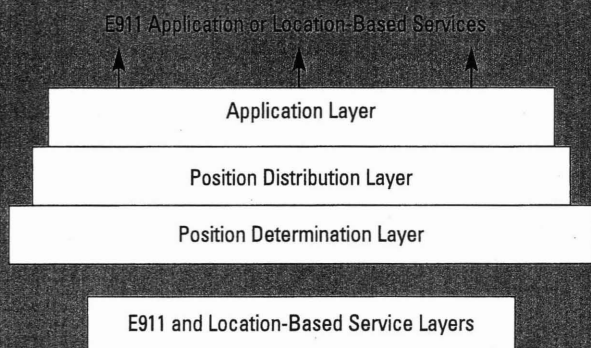
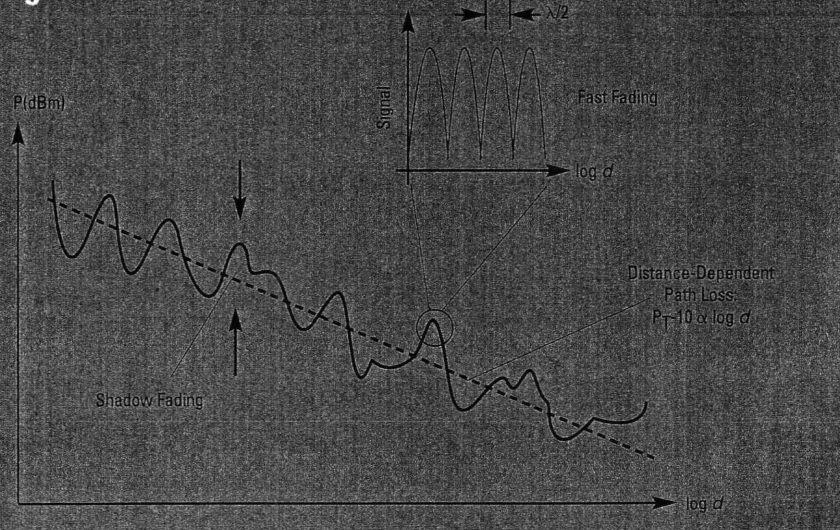


Figure 2



number. In the wireless realm, however, wireless carriers and emergency call centers must deal with the freedom of mobility—the caller can be almost

plore the technologies necessary for locating the caller as well as speeding the delivery of information to the appropriate public-safety answering point

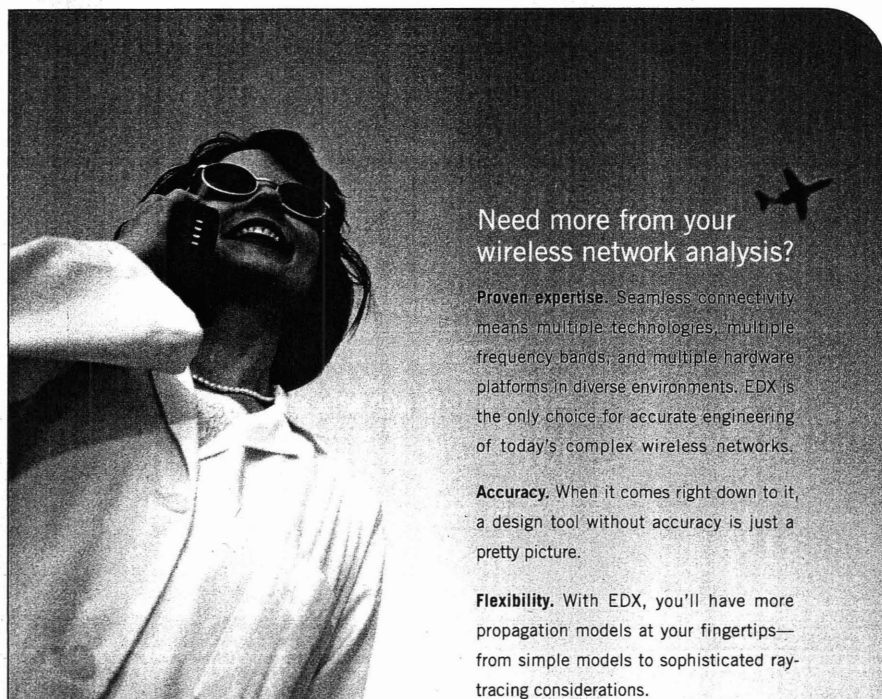
position distribution layer. This is where the location information is routed to the correct place, in this case the emergency service entity at the PSAP. Finally, the location information is handed over to the application layer for display and processing at the PSAP.

At the position determination layer, network-based or handset-based technologies are available to resolve the problem of locating the handset. The aforementioned WLS technology cap-

italizes on the signaling information already reported in the wireless network. In normal operations, handsets are continuously monitoring and reporting measurements of network conditions, such as signal strengths of cell sites, as part of the mobile handover process. These measurements create an RF "signature" that is unique for the handset's location (Figure 2). The radio signature results from unique characteristics of the path loss and shadow

fading associated with multiple channels at each X-Y location. The averaging in the measurement process removes the fast fading components.

Using sophisticated statistical pattern-matching algorithms, WLS determines the location of the user based on a geographical predicted signature database (PSD). The prediction database is built from advanced propagation models tuned with calibration measurements performed at the time of deployment, assigning strengths to each channel at an X-Y grid square (Figure 3). In GSM, the mobile handover measurements—called network measurement reports (NMRs)—are sent to the radio network about every half-second



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During an E911 call, these NMRs are aggregated to best estimate the caller's location by correlating against the PSD.

throughout a call. During an E911 call, a number of these NMRs are aggregated to best estimate the caller's location by correlating against the PSD.

Figure 4 shows a conceptual, high-level diagram of the E911 system. The main components of WLS are the following: the PSD used for the pattern-matching to determine location, the location engine that performs the real-time position estimation calculations, and the real-time network interface needed to gather the NMRs and report the position. The interfaces to the PSAP and the carrier's radio network equipment are handled through the MLC. Software automatically maintains the PSD when the wireless carrier makes network changes, such as frequency plan modifications. Operation, administration, maintenance, and provisioning (OAMP) informa-

tion for the WLS system is available through Web interfaces at the carrier's network operations center (NOC).

A detailed diagram of the E911 network architecture is illustrated in Figure 5. The communications links between GSM elements are labeled using standard nomenclature from the 3GPP specifications. The heart of the position determination layer is the serving mobile location center (SMLC), where the location engine software resides to perform the calculation of the caller's position. The SMLC pulls the real-time NMR information from the base station controller

(BSC) using an IP interface to the Abis probe units (APUs) and the Abis control function (ACF). Messaging with the BSC to coordinate the location process is performed on the 3GPP standard Lb interface over Signaling System 7 (SS7). On the back end, the SMLC has several IP interfaces to the NOC for database administration, network management and performance monitoring. The centerpiece of the position distribution layer is the gateway mobile location center (GMLC) which, working through the carrier's mobile switching center (MSC), handles the interfaces to the PSAP. The Emergency

Service Network Entity (ESNE) handles the voice call path and the Emergency Service Management Entity (ESME) performs the necessary messaging to determine the location. The ESME is where the transfer to the application layer processing occurs, to display the location information for dispatchers. Once again, standard GSM interfaces are employed between the GMLC, SMLC, ESNE, ESME and MSC.

Now that we've identified the key elements and interfaces involved in E911 location determination, we can walk through what happens during an actual emergency call. It starts with the wireless subscriber initiating a 911 call. The BSC forwards the call request to the MSC, where a voice path is set up through the ESNE to the PSAP. At the same time, the call's NMR information is cached at the ACF for use in the position determination layer. Having identified this as an emergency call, the MSC requests that the BSC geo-locate the caller; the BSC then forwards the location request to the SMLC.

Once that happens, the SMLC queries the ACF to get both the cached and new NMRs. With the NMRs, the location engine in the SMLC calculates the caller's location. This triggers the SMLC to send a response to the BSC containing the caller's estimated location. The BSC then forwards the location to the MSC, which in turn forwards the position estimate to the GMLC. The ESME queries the GMLC for the location data and the GMLC responds with the position estimate. The ESME then transfers the location to the application software for display on the PSAP terminal. Finally, the PSAP communicates the caller's location to appropriate public-safety personnel.

During an emergency call, all of these network elements must work together seamlessly to speed the location information to the dispatcher and ultimately to the first responders. We've covered the wireless E911 process from the ground up. Let's recap: Starting with the position determination layer, we looked at how the wireless caller's

Figure 3

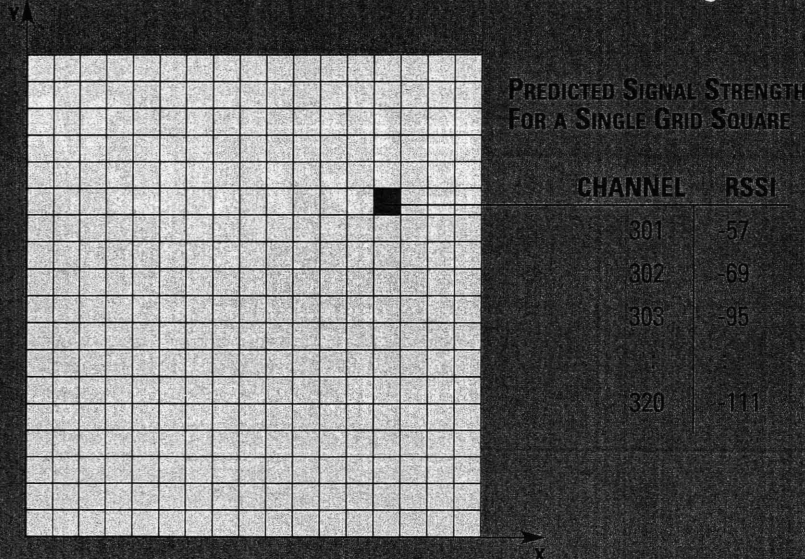


Figure 4

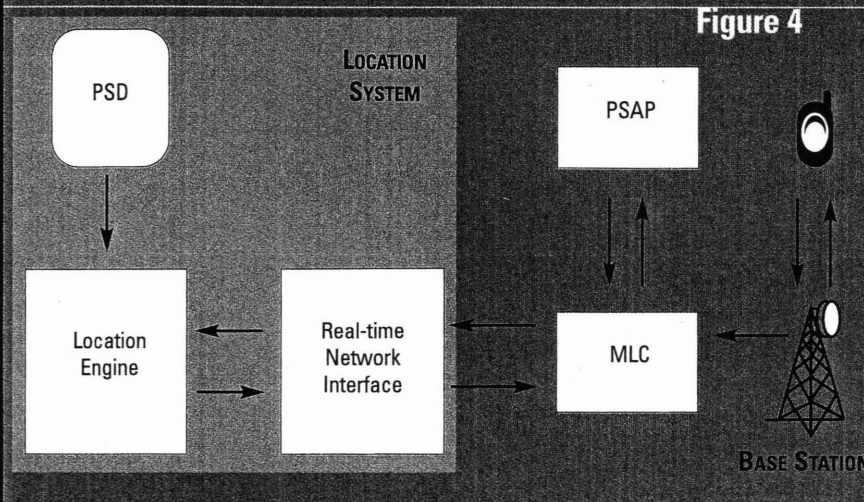
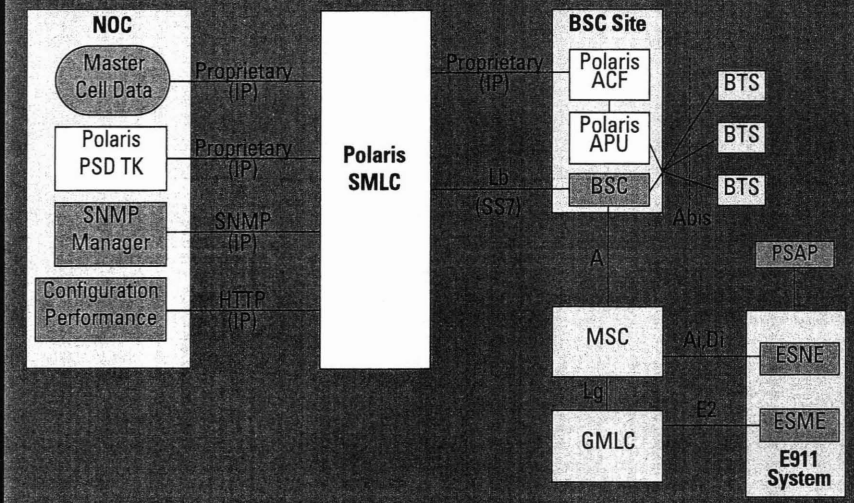


Figure 5



position is estimated using WLS technology. Sophisticated pattern-matching algorithms calculate the caller's location based on a unique signature reported by the handset. At the position distribution layer, the location in-

formation is routed through the carrier's network to the appropriate public-safety entity. A gateway serves as the interface between the carrier and the PSAP for transfer of the position information. At the PSAP, the applica-

tion layer formats and displays the location information on the dispatcher's terminal screen. The dispatcher then communicates the location to the responding emergency crews, completing the final step in the process. Armed with the E911 caller's location, the first responders are far better prepared to deal with the emergency. And that's what E911 ultimately is all about—being ready to respond. ■

Marty Feuerstein is chief technical officer for Polaris Wireless, where he leads research and development of position-location products. He has more than 20 years experience in telecom, including positions with Lucent, Metawave, U S West/AirTouch and Nortel Networks. Feuerstein also has consulted extensively in the wireless industry for companies such as T-Mobile and Ericsson, as well as a number of start-ups. He has a PhD in electrical engineering from Virginia Tech.



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Troubleshooting 101: Part II

By Harold Kinley, C.E.T.

Troubleshooting is the science (or art?) of finding the defective section, stage and component of a malfunctioning piece of equipment in the shortest possible time. Almost anyone can replace a defective component, but finding the culprit in a complex circuit is what separates the men from the boys in the shop!

Many years ago, I was doing come CB repair work on a moonlighting basis. A gentleman brought in a CB transceiver, and although I was busy at the time, he convinced me to drop what I was doing to work on his defective transceiver. After about 30 minutes of troubleshooting work, I found the culprit to be a leaky

capacitor. After I showed the customer the problem, he asked how much the repair would cost. After I informed him that the repair charge would be \$20, he advised me to "button up" the radio and not to repair it.

So, I "buttoned up" the radio and charged the gentleman(?) \$15 for the troubleshooting time. I knew he would simply buy a replacement capacitor and replace it himself after I had identified the defective component for him. He had some ability with a soldering iron but not much at troubleshooting. Again, anyone can *replace* a component, but it takes a real troubleshooter to *find* the defective component(s).

This month we will look at a couple of troubleshooting techniques that, al-

though a little unorthodox, are quite useful.

Signal tracing

Suppose you are troubleshooting a receiver that has a weak (low gain) or dead stage. You can force a high-level signal through the receiver, but the sensitivity is poor. First, the defective stage must be located.

In AM receivers, a simple demodulator (demod) probe can be used in conjunction with an audio amplifier to trace the AM signal through the receiver. Aside from the limiter stages and detector, the FM receiver is no different from the AM receiver. This means that a simple demodulator probe can be used to trace the signal in the FM receiver.

Besides the demodulator probe, an audio amplifier with optional metering is needed. The audio amplifier simply amplifies the *demod* probe output and allows it to be heard in the loudspeaker. To use this method, simply apply a high-level signal to the input of the receiver. The signal must be *amplitude-modulated* at a high level (about 70% to 80%) using a 1kHz tone. By moving the demod probe from input to output of the various stages, the malfunctioning stage can be located. Relative stage gain is indicated by changes in the tone level from input to output.

If you happen to have a SINAD meter with a built-in audio amplifier and a voltage scale with decibel calibrations, so much the better. The demod probe can be connected to the input of the SINAD meter, and the voltmeter can be used as a better indication of stage gain, while the speaker serves as aural indication (if you are simply signal-tracing and not interested in stage-gain measurements).

It is important to keep the input signal level as low as possible to prevent overloading the audio amplifier or the SINAD meter. The preferred method is to set the RF signal input level just high enough to produce a usable reading at the SINAD meter-audio amplifier with the demod

(continued on page 50)

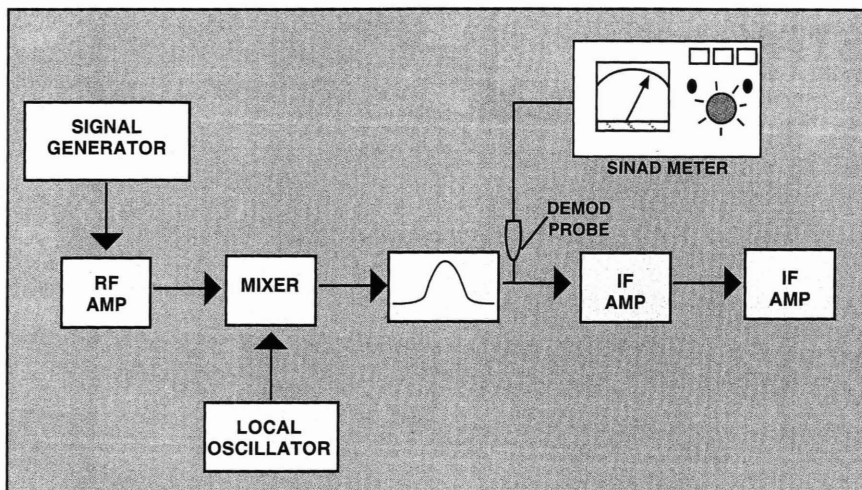


Figure 1. To use a SINAD meter as a signal tracer, just hook the demod probe to the SINAD meter input.

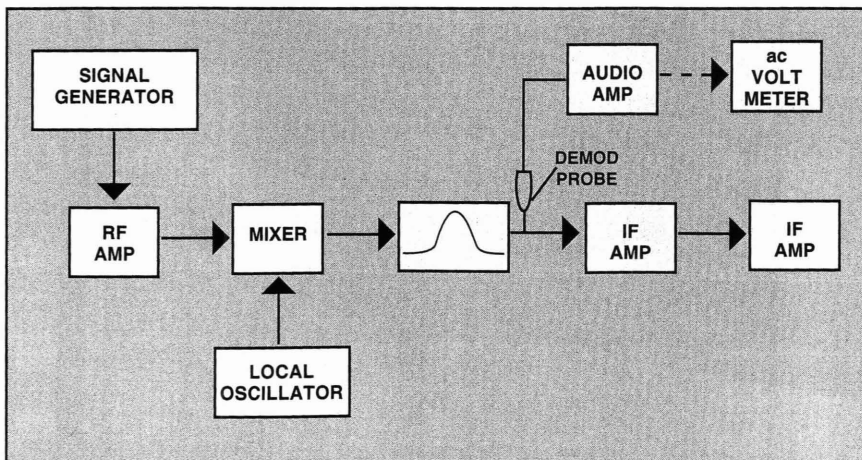


Figure 2. An audio amplifier, such as the Radio Shack model 277-1008, can be used as a signal tracer with a demod probe as shown. An ac voltmeter can be plugged into the phone jack to get a voltage reading as a reference.

Kinley, a certified electronics technician, is regional communications manager, South Carolina Forestry Commission, Spartanburg, SC. He is a member of the Radio Club of America. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, which is available for direct purchase. Write to 204 Tanglewyld Drive, Spartanburg, SC 29301.

Technically speaking

two capacitors and a diode connected between the voltage rail and ground. One of the components is shorted, but which one? Each component could be disconnected one at a time to observe the effect on current drain, but suppose they are very difficult to gain access to—what then?

Since printed circuit foils possess a certain amount of resistance (depending on thickness, width and length) a small amount of voltage drop will appear across the foil when current flows through it. That is the basis for the following procedure.

A digital voltmeter is connected across the ends of the voltage rail as shown in Figure 3B on page 50. Then a battery is connected, as shown, with a 2Ω resistor connected in series. The resistor serves to limit current flow through the circuit foil in order to protect the foil from overheating and burning. At the same time, enough current is allowed to flow so that a measurable voltage drop is produced. At Figure 3B, let's suppose that capacitor C1 is shorted. With the resistor connected at the point shown, the dc voltmeter will indicate a *negative* reading.

That means the probe has been placed on the negative side of the short as referenced to the voltmeter terminals.

If the probe is moved farther to the right as shown in Figure 3C on page 50, then the voltmeter will indicate a more negative value, indicating that the probe is farther removed from the short circuit in the direction of the negative voltmeter terminal.

If the probe is repositioned as shown in Figure 3D on page 50, then the voltmeter will indicate a *positive* reading, meaning that the probe has been placed on the positive side of the short as referenced to the voltmeter terminals. If the probe is moved farther to the left (Figure 3E on page 50), then the voltmeter will indicate a more positive value, indicating that the probe is farther removed from the short circuit in the direction of the positive voltmeter terminal.

To locate the short circuit, the probe is moved along the rail until the voltmeter reads "000." This is the pinpoint location of the short circuit, and the component connected to this point is the culprit.

Instead of using the battery with the

current-limiting resistor, a constant-current power supply could have been used. Be sure to limit the current to the minimum amount that provides a usable reading on the voltmeter. In some cases, the current may have to be increased to produce a sufficient voltage drop for this method to work. Just be sure to limit the amount of time the probe is connected to prevent burning the circuit trace.

Another useful tool for locating short circuits is the "Shortsqueek" from Global Specialties. It emits an audible tone, and as the probe is moved closer to the location of the short circuit, the tone increases in frequency. By listening to the tone frequency you can locate the source of the short circuit.

Whether you are troubleshooting down to the component or to the board, don't be afraid to try new troubleshooting techniques. Some techniques are more efficient than others, so try to use the technique and equipment that fits the situation. Learn to adapt.

Next time, we will conclude this mini-series on troubleshooting.

So, until next time—*stay tuned!*



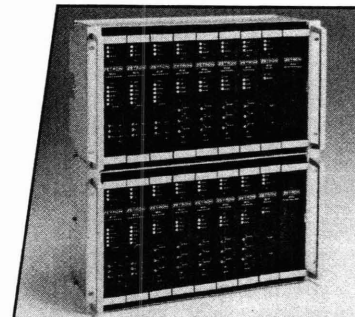
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Troubleshooting 101: Part III

By Harold Kinley, C.E.T.

The ability to troubleshoot electronic circuitry effectively and efficiently is composed of three things: knowledge, experience and an innate ability. The reference to an *innate* ability may sound unscientific and even ludicrous, but ask the old veteran service-shop managers who have "been through" many technicians over the years, and they will tell you that there is something besides knowledge and experience that goes into making the expert troubleshooter. It is intangible, but it marks the true troubleshooter!

Notice that I haven't mentioned test equipment? Test equipment is essential to troubleshooting complex electronic equipment, but it does not a troubleshooter make. Give your worst troubleshooter the best equipment and your best troubleshooter the worst equipment, and who do you think will get the job done? This is not to say that good test equipment is not necessary for good troubleshooters. It has to do with efficiency and doing the best possible job. Good test equipment in the hands of a good troubleshooter will yield the best possible results in the shortest possible time.

Resources

When troubleshooting a piece of equipment, use all the resources that may be at your disposal. This is not limited to test equipment. Don't overlook technical acquaintances or associate technicians who might have successfully dealt with the same problem in the past. Human resources are usually the best resources.

If you have another piece of equipment identical to the equipment you are troubleshooting, use it as a reference in analyzing test and measurement procedures you are performing on the ailing equipment. This expedites repairs and avoids a wild goose chase!

Keep notes on unusual problems and their solutions for future reference. Share them with your fellow technicians. Not only will they think you are a swell per-

son, they might even buy your lunch! As we get older, the notes are even more important. Ever run across a strange problem and get a feeling of *déjà vu*? Yep—been there—done that—just can't remember the solution, only the problem! Now, you have to reinvent the wheel. If *only* you had made a note—or kept the note you did make where you could find it! Oh well, "next time," will you say?

Other resources are the troubleshooting flow charts that are commonly found in service manuals. These insult your intelligence, you say? Well, I used to think the same way until I really put one of these charts to the test on an occasion. Some of

often the best place to start. This gives you a quick and logical overview of how all the individual circuit boards or subsystems are integrated to form the complete system.

More troubleshooting techniques

There are many different troubleshooting techniques. Basically, they can be reduced to a few when generalized as *voltage-resistance-current measurements*, *waveform checks*, *signal injection* or *signal tracing*. Actually, signal tracing and signal injection are about the same—one is simply the complement of the other.

Technically speaking, signal injection and signal tracing must be used together, since there must be a signal injected somewhere to be traced, and vice versa.

When testing mobile radio receivers on the bench, be sure to check the operation of the transceiver over a voltage range of 11V–15V. Even higher (or lower) voltages may occur in mobile installations, but the transceiver cannot be expected to operate properly at extremely high ends of the voltage range. Certain problems in particular transceiver models tend to show up on a predictable basis at high or low supply voltages.

One case history in particular comes to mind. A firefighter complained that his transceiver (installed in a diesel crawler tractor) started "howling" while in use on a fire suppression

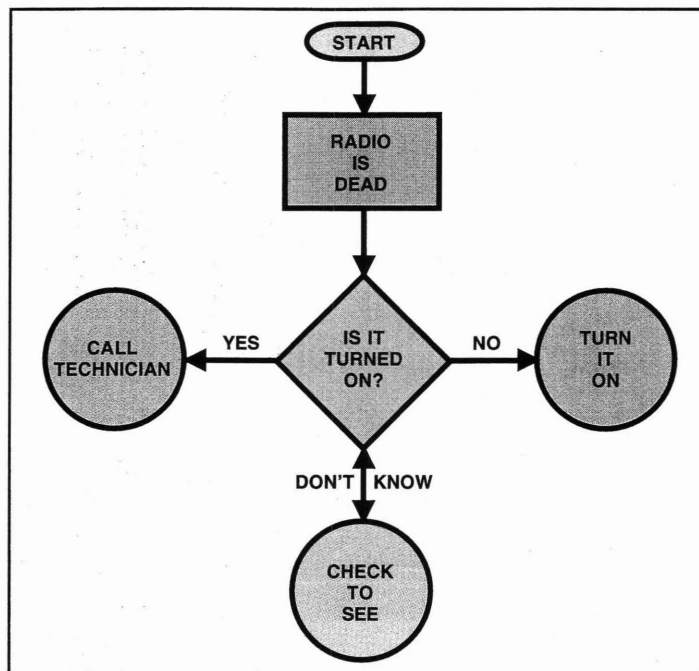


Figure 1. Some service manual troubleshooting flow charts like this one may seem to insult your intelligence, but good charts can be a valuable resource.

these charts really work, given the chance. Granted, some are much better than others. An example of a *bad* flow chart is shown in Figure 1 above. This is a deliberate exaggeration, but not too far from some I have seen!

When all else fails, *study the manual!* You may have to do some in-depth reading and hard concentration. The solution to your problem may not appear in the manual in so many words, but often something in the text will trigger a thought process that leads you right to the trouble. Studying the manual includes studying the schematic diagram as well as the functional block diagram. In complex equipment, the functional block diagram is

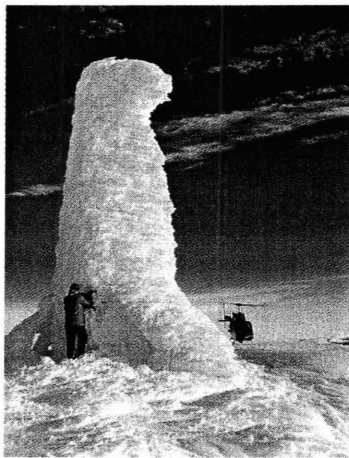
incident.

To gain more insight into the nature of the trouble, I carefully quizzed the firefighter. I learned that the volume control setting had no effect on the "howling," which would begin after the tractor

(continued on page 52)

Kinley, a certified electronics technician, is regional communications manager, South Carolina Forestry Commission, Spartanburg, SC. He is a member of the Radio Club of America. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, which is available for direct purchase. Write to 204 Tanglewyld Drive, Spartanburg, SC 29301.

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Technically speaking

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got "warmed up" and would only cease when the tractor was throttled down and the radio was turned off and then back on.

I had the firefighter bring the tractor into the installation bay next to the radio shop so that I could begin troubleshooting with the radio still installed in the tractor.

With the tractor not running, I turned the radio on and found it to be operating normally. Then I started the tractor and let the engine run at low throttle for several minutes—still no evidence of trouble from the radio. Then I increased the throttle until it reached a point where the radio started "howling." Yes, indeed, I *do* mean "howling!"

As the firefighter had stated, the volume control had no effect on the howling. Next, I throttled down the tractor engine—but the howling persisted. I turned the radio off and back on, and the howling was gone. Now, what was the cause?

The next step was to connect a dc voltmeter across the battery and repeat the previous process. The initial battery voltage was around 13V, but when the engine was started and throttled up, the battery voltage started climbing. When the battery voltage reached the 16V threshold, the radio would start howling. Aha! The voltage regulator! The regulator was replaced and that took care of the problem.

Since then, I have encountered the same problem many times, and the voltage regulator has always been the culprit.

Again, check mobile equipment over a range of supply voltages in verifying complaints or in troubleshooting a problem.

In modern transceiver circuits, you will usually find several different supply voltages from the power supply or power distribution board. The digital circuitry generally operates from a 5V supply point. There may be other supply voltages as well (including the full battery voltage for mobile transceivers).

For base stations, the rule is the same—check all supply voltages. It is also a good practice to scope the supply voltage points to determine if the supply voltage is clean. The ripple component should be very low and free of spikes. Also, for regulated supply points, check to make sure the voltage regulation is within bounds.

Don't overlook using the spectrum analyzer as a signal-tracing tool when troubleshooting receivers and exciters. Watch the maximum input level to avoid damaging the front end of the spectrum analyzer. If the input impedance of the spectrum analyzer is 50Ω, then you can expect it to cause serious loading of the circuit under test. However, you can use

a simple and inexpensive low-capacitance scope probe (10/1) to reduce the loading effects. Any attenuation caused can be reduced by the use of a preamplifier between the scope probe and the analyzer input.

This procedure works particularly well if you use a known good transceiver to establish reference measurement levels ahead of time using the same test equipment, including probes and preamplifier. This establishes a good benchmark for future reference and is particularly valuable if you service many transceivers of the same model. There are manifold returns on the time used in making such measurements.

Many manufacturers' spectrum analyzers also offer a special preamplified, high-impedance, signal-tracing probe that allows the analyzer to be used for signal tracing in circuits with minimal loading effects. Such probes are called *active* probes. The probe supply voltage usually comes directly from the spectrum analyzer.

Conclusion

The best advice in troubleshooting is to *keep an open mind*. Let your troubleshooting tests and measurements lead you to the problem. Don't guess what you think the problem may be and then set out with test procedures to prove that you are right.

No test procedure is useless. Even though it may be way off base, it will provide you with some information. However, each test procedure should lead you one step closer to the problem part or board. Don't rule out any component based on appearance alone. No matter how innocuous it may look, it *can* be bad! Electronic components often look like new on the outside, but are totally "zapped" on the inside!

Make sure you thoroughly understand the circuit you are troubleshooting. Without an understanding of how the circuit *works*, you will not be able to understand what makes it *not* work.

Know and understand the limitations of your equipment and the situations where false readings can occur. Know when to get a second opinion by another method.

Keep your test equipment properly maintained and calibrated. Keep fresh cells in battery-operated equipment and have spare batteries on hand. Keep your test leads and probes in good condition.

By following some of the basics set forth in this series, along with hard work, you can become a better troubleshooter. By successfully meeting the challenge of each troubleshooting job, you gain experience, knowledge and confidence that will allow you to successfully meet greater challenges.

Until next time... *stay tuned!*



Locating cable faults with the time domain reflectometer

By Harold Kinley, C.E.T.

Opens, shorts or less-severe impedance discontinuities have a way of showing up on cables in strange places—places you might never suspect. These can occur on coaxial transmission lines or twisted-pair lines. Such opens, shorts or other impedance discontinuities are called *faults*. The location of faults cannot be determined with simple ohmmeters. Even the existence of certain faults cannot be determined with an ohmmeter. After all, we can't begin to correct a fault unless we know where it is or at least have some

idea where to start looking for it.

Time domain reflectometer

The basic principle of the time domain reflectometer (TDR) is fairly simple. See Figure 1 below. A pulse is generated and sent down the line. If the line is terminated in a load impedance that is equal to the characteristic impedance of the line ($Z_0 = Z_L$), no reflected wave or echo will be returning to the generator or source. Let P_i represent the incident pulse generated by the pulse generator and P_r represent the echo or pulse reflected by the impedance discontinuity. By measuring the time difference between the incident

pulse (P_i) and the echo or reflected pulse (P_r), it is possible to determine the distance to the fault that produced the echo pulse. Calculations must include the velocity factor of the cable.

If a transmission cable has a velocity factor of 1.0, then a pulse will propagate down the line at 0.982 foot per nanosecond (0.982ft/ns). If the velocity factor of a transmission cable is 0.66 and the time difference between P_i and P_r is 430ns as measured on the time base of the oscilloscope, then the distance traveled by the echo pulse is: $0.66 \times 0.982 \times 430 = 278.69$. Thus the echo pulse has traveled 278.69 feet in 430ns. To get the distance from the source to the fault, we must divide this figure (278.69) by 2 to get 139.35 feet, the distance that the echo must travel back up the line to reach the point of measurement on the scope. The general formula for this equation is:

$$D = \frac{0.982VT}{2}$$

where

D = distance to fault in feet

V = velocity factor

T = time between pulses in nanoseconds.

Cablemate Analyst reflectometer

The Cablemate Analyst reflectometer from Advanced Electronic Applications (a division of Tempo Research, Vista, CA) is a low-priced, feature-packed TDR in a small, hand-held unit. All of the features of this unit cannot be described in this space; however, some of the more essential features will be presented. Although it is not necessary, in most cases, to use an oscilloscope with this instrument, the manufacturer has made it convenient to do so. The use of an oscilloscope makes it possible to see minor reflections that might not be picked up by the instrument.

Using the oscilloscope

The oscilloscope is connected to the reflectometer exactly as shown in Figure 1.

(continued on page 38)

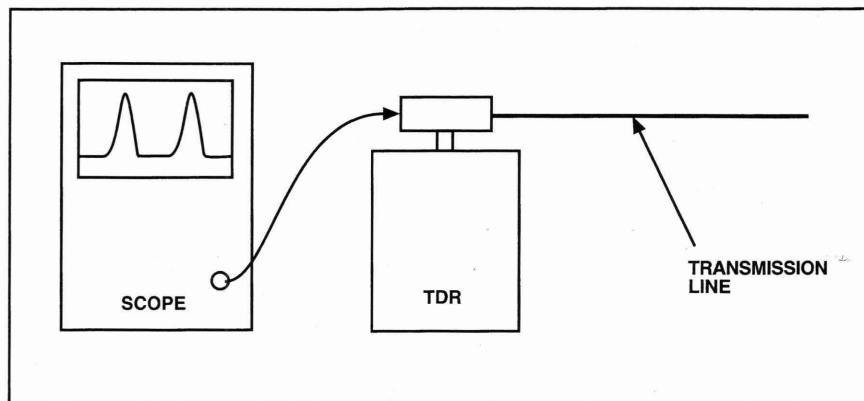


Figure 1. To closely examine the echo pulse(s), an oscilloscope is connected to the TDR. By measuring the time delay between the incident and echo pulses, the distance to the fault can be determined. The Cablemate Analyst reflectometer makes it simple to perform this calculation.

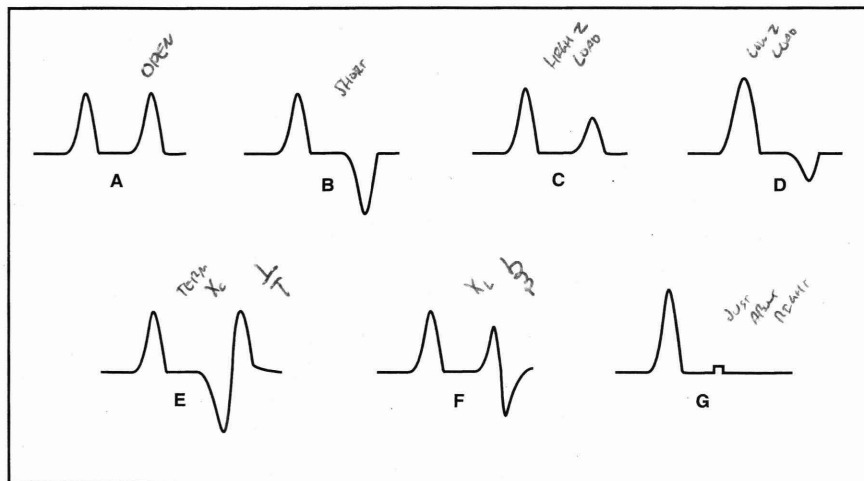


Figure 2. Waveforms for various load impedances. At A the transmission line is open, resulting in a positive echo pulse nearly equal to the incident pulse. At B the transmission line is shorted, resulting in a negative echo pulse nearly equal to the incident pulse. At C the transmission line is terminated in a load impedance much higher than the characteristic impedance of the line. At D the transmission line is terminated in a load impedance much less than the characteristic impedance of the line. At E the transmission line is terminated in a capacitive reactance. At F the transmission line is terminated in an inductive reactance. At G the transmission line is terminated in a load impedance slightly greater than the characteristic impedance of the line, resulting in a minor reflection.

Kinley, a certified electronics technician, is regional communications manager, South Carolina Forestry Commission, Spartanburg, SC. He is a member of the Radio Club of America. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, which is available for direct purchase. Write to 204 Tanglewyld Drive, Spartanburg, SC 29301. Kinley's email address is hkinley@aol.com.

Technically speaking

(continued from page 8)

When using coax cable, a BNC Tee connector is needed. The oscilloscope connects to one side of the Tee connector and the transmission line to the other side. When twisted pair cable is used, the twisted pair connects to the binding posts, and the oscilloscope to the BNC jack. In this case a 50Ω cable is connected to the reflectometer. Figure 2 on page 8 shows

several waveforms for various load impedances. The TDR will provide the distance to the fault if you enter the time delay between the incident and echo pulse. For example, if you examine the timebase of the scope and determine that the time between the incident pulse and the echo pulse is 400ns, just enter the number 400 into the reflectometer, and it will return the distance of 130 feet. If you use an oscilloscope, make sure that the

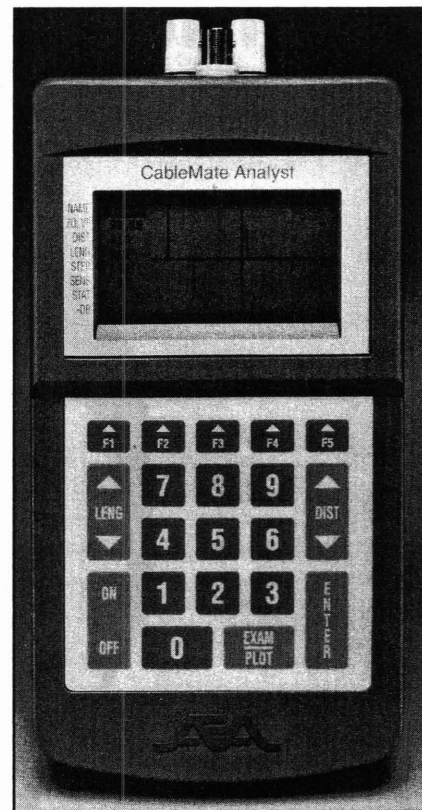


Photo 1. A time domain reflectometer helps locate cable faults that may show up in strange places.

time base is accurate. Otherwise, your distance calculations will be inaccurate. Usually, we are looking for the distance to a major fault—open or short circuit. For this application an oscilloscope is not needed. The oscilloscope is only needed for locating minor reflections or for observing the echo waveforms.

Selecting the transmission line

As many as 96 transmission line types can be stored in the reflectometer for instant recall. The unit comes preloaded with many lines. You can add transmission lines by setting up the characteristic impedance, velocity factor and then storing under a name you choose. A BNC jack is used for connecting coaxial cables, and binding posts are used for connecting twisted-pair lines. The binding posts apparently connect through a balun that provides a 4:1 impedance change. To test a 600Ω balanced line, the line would connect to the binding posts, and the impedance would be set to 150Ω so that when passing through the balun, the impedance would be transformed to 600Ω.

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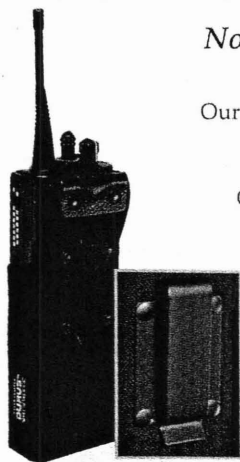
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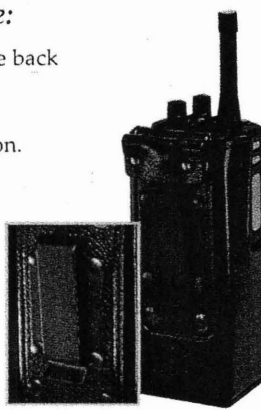
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along with other vital information. The echo is plotted, and a cursor is provided to enable an examination of the echo(s). Upon placing the cursor over the echo, the information is provided on the screen. Information such as distance-to-fault and return loss of echo (an indication of the severity of the mismatch) is provided.

General information

The reflectometer can locate faults from

16 feet to 2,000 feet (5 meters to 600 meters). Faults less than 10 feet apart can be resolved on the display. Accuracy is rated at $\pm(1\%+2')$. For example, at 200 feet the accuracy would be $\pm 4'$. Characteristic impedance for coaxial cable is 45Ω to 150Ω in 7Ω steps. For balance line (two-wire cable), the characteristic impedance range is 180Ω to 600Ω in 28Ω steps. The velocity factor range is 30% to 99% in 1% steps. The output pulse is +5V with a pulse

duration of 25ns. The unit can run with supplied ac power supply or a battery pack (8 AA cells). Battery life is about 12 to 14 hours. Cable types and plots along with modes, settings and calibration data are stored in non-volatile memory. A serial port is provided for connection to a computer for use with optional software. The unit weighs slightly over 1.5 pounds including batteries.

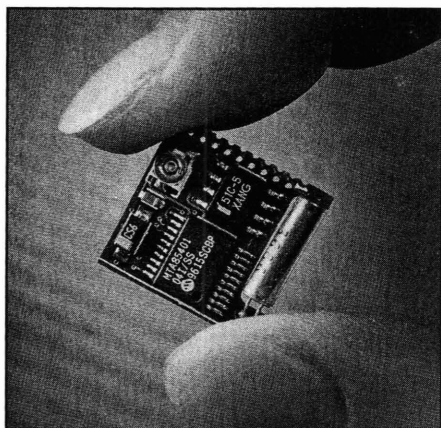
Summary

The CableMate TDR is user-friendly. In an hour or so one can become sufficiently familiar with the unit to make useful measurements with it. The menu is easy to understand, and condensed instructions are printed on the back of the unit. Such an instrument is useful for making reference measurements on transmission lines for future maintenance reference, as well as for determining line length, velocity factor and fault locations. For further information on this instrument, contact Advanced Electronic Applications (Division of Tempo Research) 1221 Liberty Way, Vista, CA 92803. Phone: 760-598-9677.

Until next time—stay tuned!



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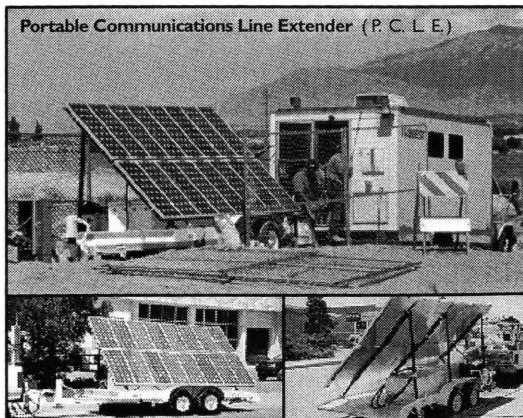
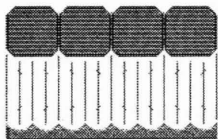
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Analog multiplex systems: The basics

Part 1 — Multiplexing combines multiple discrete signals into one composite transmission. Familiarity with direct-to-line systems enhances a technician's repertoire.

By Jeff Ashley

A common complaint for employers in today's communications industry is not being able to find *qualified* technicians. Years ago, two-way field technicians might have been considered "qualified" by merely having experience with Micors or Master IIs—not so today. Employers are looking for people whose experience extends to microwave and multiplex systems. Those seeking employment must be willing to expand their knowledge base if they expect to be marketable. Two-way technicians who are *currently* employed must be willing to expand their responsibilities to maintain job security.

With this in mind, let's take a brief and practical look at the subject of analog mul-

Ashley is a communications technician in Los Angeles.

tiplex systems, which can prove valuable to those yet unfamiliar with such transmission systems equipment.

Analog mux—FDM

Frequency-division multiplexing (FDM) is, and has been for quite some time, a common analog method of combining or encoding multiple discrete signals into one composite transmission signal. Analog microwave radio systems are the typical transport mechanisms for FDM systems. The multiplex (mux) equipment provides the means for interfacing the individual circuits (radio, voice or data) with the microwave radio. For every circuit transported over the microwave system, a corresponding mux channel modem will be on each end of the circuit.

In general, there are two types of FDM systems: *basic* and *direct-to-line*. Basic FDM systems preceded the arrival of direct-to-line equipment by decades. The ar-

chitecture of these older systems typically requires multiple equipment racks to accommodate all the individual, interconnected units necessary for each step of the multiplex process.

Over the years, advances in technology and miniaturization of circuitry made possible the direct-to-line method that offers the same end result with significantly less equipment. Direct-to-line systems allow companies with minimal channel requirements to obtain multiplex equipment on a per-channel basis rather than as a full-blown system made to accommodate hundreds of channels, whether you need them all or not.

Because of their cost-effectiveness, direct-to-line systems are probably what today's communications technician will encounter. For this reason, basic FDM systems will only be mentioned briefly as a backdrop to the more modern and popular direct-to-line systems.

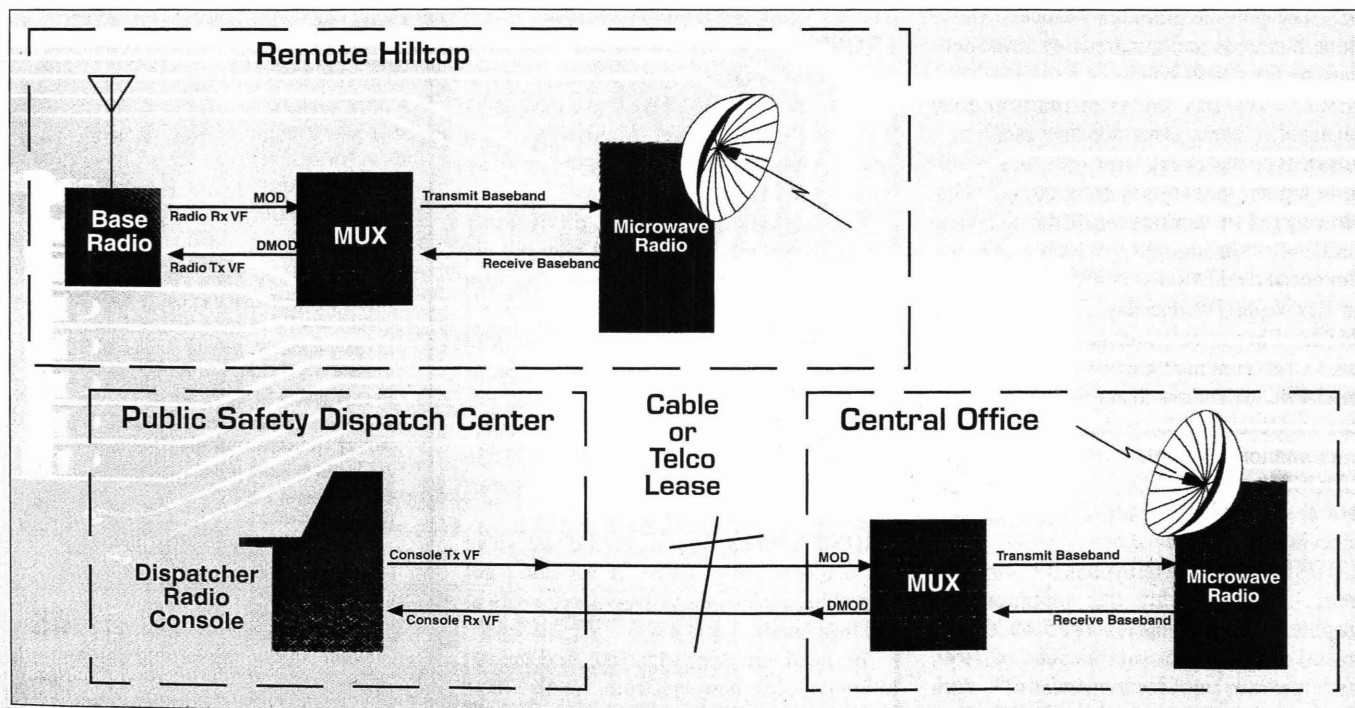


Figure 1. Possible remote-controlled radio circuit.

VF and signaling

Multiplex and microwave systems are often referred to as *transmission systems* equipment because they provide a method of transmitting broadband signals from one point to another. A common term used in the transmission systems field is "VF," short for "voice frequency." A VF signal is the audio information being carried over a circuit. The VF transmit connection to the mux channel modem is referred to as the *MOD* while the receive connection is the *DMOD*. (See Figure 1 on page 72.)

In addition to VF signals, the mux circuit often carries information called *signaling*, which relates supervisory information from one end of the circuit to the other. Using a subscriber telephone circuit as an example, a method is needed to signal the central office as to whether the phone is on- or off-hook, or to activate ringing. Radio circuit signaling is used to convey the push-to-talk (PTT) function from the dispatcher console to the remote base station. VF and signaling can be accomplished over the same multiplex channel at the same time.

The primary mux modem signaling connections are referred to as the *M-lead* (transmit) and *E-lead* (receive). Although modem-strapping options usually accommodate alternate arrangements, transmit signaling is typically activated by applying battery (the modem's operating voltage) to the M-lead connection. When signaling is detected from the far-end mux modem, the local E-lead is activated, which most often applies a ground to its external connection. As with the M-lead, onboard E-lead strapping options usually allow for alternatives such as the internal battery voltage to be sent

out to the external connection. In some cases, mux modems are equipped with an *F-lead* that allows some other external voltage level to be applied to the modem, which can then be gated out the E-lead when signaling is detected. This is an advantage when the external device activated by the E-lead requires a specific voltage that is not obtainable from the mux modem itself.

VF frequency response

The frequency response of speech can range from about 100Hz to 8kHz. However, the primary components necessary for speech recognition are contained below 3kHz. The typical analog mux channel has a circuit frequency response of 300Hz to 3,400Hz. Frequencies above and below this range are rolled off with filters.

FDM—direct-to-line method

In direct-to-line systems, the whole multiplexing process takes place in one unit, the mux channel modem. (See Figure 2 below.)

Signals are transmitted to a far-end location by applying analog circuit information (VF) to the MOD of the mux channel. The modem processes the signal through a limiter stage to prevent excessive levels from overdriving the following circuits. The VF signals are then routed through a high-and-low-pass filter to give the desired 300Hz to 3,400Hz frequency response. A VF amplifier stage follows and allows adjustment of the transmit VF levels. The VF signal is applied to one port of the first balanced modulator while an IF oscillator (typically in the 3MHz to 6MHz range) is injected into another. The resultant mix translates the original 300Hz-to-3,400Hz VF signal into the RF

spectrum as a double-sideband, suppressed-carrier (DSBSC) signal. A crystal filter then eliminates the unwanted sideband, leaving a single-sideband, suppressed carrier (SSBSC) signal occupying a 4kHz bandwidth.

This modulated signal is then applied to a second balanced modulator stage that again produces a DSBSC signal whose unwanted sideband and suppressed carrier are eliminated by filtering. The final result is a SSBSC signal modulated with the original VF circuit information. At this point, the signal is considered HF (high frequency), rather than VF.

Each mux circuit has the same first IF frequency. The second modulator, however, has a different and unique frequency injected for each channel in the system. This prevents one mux channel from interfering with another by ending up on the same frequency in the HF spectrum. To accommodate this, the second oscillator (which typically covers the 4MHz to 8MHz range) is a dip-switch programmable, phase-lock-loop circuit allowing frequency adjustment in 4kHz steps. Spectrally, the channels form a collection of adjacent HF frequency slots, one for each circuit. Each slot has a bandwidth of 4kHz and contains the modulated VF signals unique to that circuit. The spectral composite of all circuits in the system is referred to as the "baseband."

The mux channel modems on both ends of the same circuit must have their programmable oscillators set to the same frequency so that one end can communicate with the other. The baseband channel slots follow a CCITT-recommended scheme that establishes a standard for channel placement

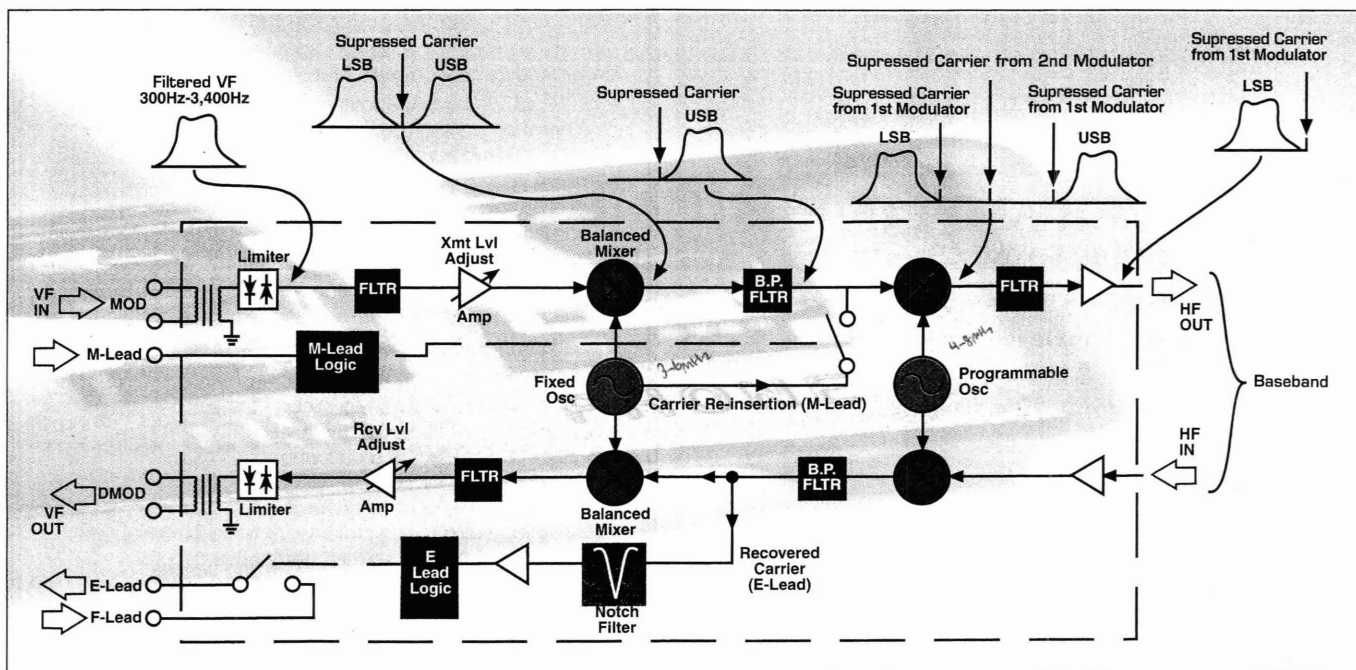


Figure 2. Direct-to-line mux channel modem.

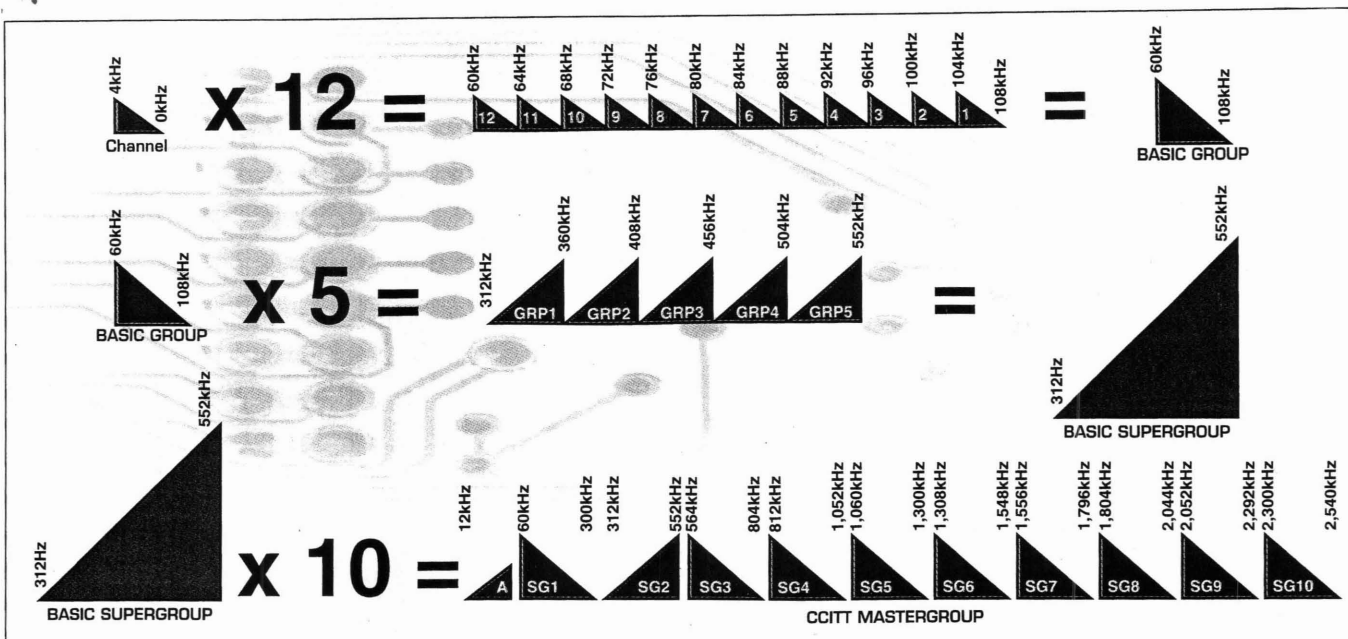


Figure 3. Basic groups.

within the baseband spectrum.

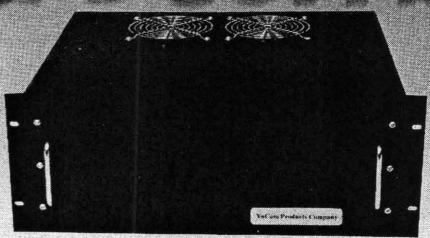
Once multiplexed to the proper frequency slots on the baseband, the circuits may be transmitted to the distant-end by means of an analog microwave radio transmitter. The mux system baseband is connected to the transmitter through a coaxial cable, which in this case would be referred to as the trans-

mit baseband cable. At the distant-end, the baseband signals are demodulated through the microwave receiver. They are then applied, via the receive baseband cable, to a complementary FDM system. Here, the HF baseband signals are de-multiplexed through a process identical to the multiplexing of the opposite end, only in reverse. The

end result is a recovered VF waveform the same as that applied to the mux channel MOD on the distant end.

Groups, supergroups and mastergroups

Basic mux systems accomplish the multiplexing process through a standardized series of intermediate steps within a variety of

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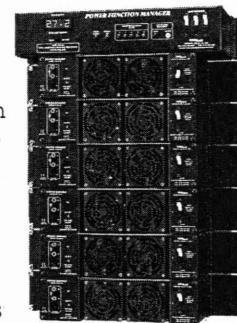
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separate, but interconnected, units.

Channel units are shelved in groups of 12. Each unit in the shelf has a different carrier frequency injected into its modulator stage by external channel carrier generator units. The result is a group of 12 channels translated one above the other with respect to frequency, occupying 48kHz of bandwidth between 60kHz and 108kHz. At this point the 12 channels are referred to as a basic group. Because each basic group occupies the exact same frequency range, they must be cabled separately to the next stage in the process to avoid interfering with one another.

These basic groups are modulated up to yet a higher frequency range, five at a time. Each group has a different injection frequency applied from its respective group carrier generator unit. The result is five groups (of 12 channels each) stacked one above the other with regard to frequency. The five groups are now referred to as a basic supergroup and occupy 240kHz of bandwidth between 312kHz and 552kHz.

Each basic supergroup occupies the same frequency spectrum and must, as with the basic group equipment, be cabled separately to the next stage.

Basic supergroups are then applied, in sets of 10, to higher-level translation equip-

ment. Different injection carrier frequencies are again supplied, this time by supergroup carrier generator units. The results are 10 supergroups stacked one above the other on the baseband, now referred to as a mastergroup.

The translation processes can continue beyond the mastergroup level. For industrial applications, however, this is typically the greatest required channel capacity (600 channels).

Each translation process produces sum and difference frequencies that correlate to upper and lower sideband signals. As seen in Figure 3 on page 74, the CCITT modulation scheme dictates upper sideband (USB) for Supergroup Two while the others are lower sideband (LSB).

Unassigned spectrum is left available below 60kHz and can be used for additional channel assignments. The 12kHz-60kHz portion of the baseband is referred to as Group A. Like Supergroup Two, Group A comprises USB channels. The 0kHz-4kHz portion of the baseband is used as a maintenance communications channel on the microwave system, called an order wire. It allows communication between technicians at different microwave sites sharing the same baseband. Two more channels can be used in the 4kHz-12kHz spectrum. Often, remote

alarm monitoring & control systems occupy one of these channel slots.

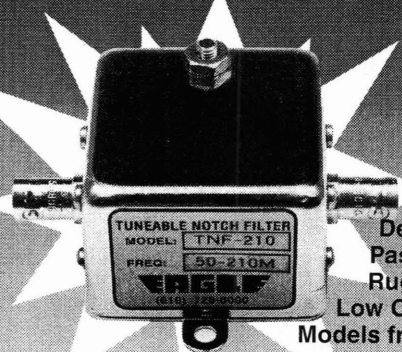
Synchronization

In both basic and direct-to-line FDM systems, pilot tones can be placed on the baseband for synchronization between the central office and the remote-end mux locations. A pilot is a specific frequency tone such as 60kHz, 64kHz, 308kHz, 512kHz or 564kHz, sent at a specific level. Once received at the remote site, the pilot tone frequency is divided down and used as a reference for the remote-end multiplexing and demultiplexing processes.

Synchronization can also be achieved through master/slave strapping on the mux channel modem. A channel modem, strapped as the slave, locks onto the master's signaling tone frequency. The master tone frequency is used within the slave channel modem for the multiplexing and demultiplexing processes.

Whichever method is used, the purpose of synchronization is to eliminate translation error between the two ends of the circuit. Excessive translation error on a voice circuit makes the party on the other end of the circuit sound like Donald Duck. Those familiar with tuning VFOs on SSB receivers will recognize the sound. ■

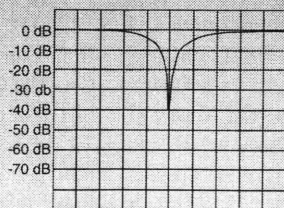
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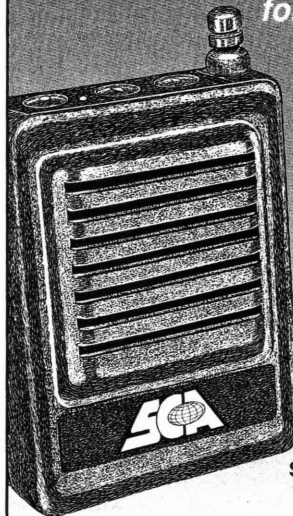
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Analog multiplex systems: The basics

Part 2 — Technicians need familiarity with modem circuitry, signaling formats and proper signal levels when servicing multiplex systems.

By Jeff Ashley

Today's two-way radio field technicians have increased responsibilities that extend to maintenance and installation of multiplex equipment. As such, they encounter terminology and a variety of circuit configurations that are foreign to the two-way industry.

Four-wire & two-wire circuits

Most communications circuits are designed to send and receive information, often at the same time. Analog multiplex (mux) systems accomplish this full-duplex operation through use of a four-wire (4W) mux modem. The modem has separate transmit and receive circuitry, providing isolation between the two. The term 4W comes from the number of VF wire connections made to the mux channel modem. Because the MOD and DMOD are balanced connections, each requires two wires, totaling four, hence the term 4W. (See Figure 1 below.)

Ashley is a communications technician in Los Angeles.

In the case of a telephone instrument or other two-wire (2W) device, the transmit and receive signals are carried on a single, balanced pair of wires. A 2W circuit must be converted to 4W when being applied to a transport system having separate receive and transmit circuitry. This conversion is accomplished through a 2W/4W hybrid circuit that is a three-port transformer arrangement. The telephone connects to the 2W port, while the mux modem connects to the 4W receive and transmit ports.

Signaling—in-band & out-of-band

Circuits usually require supervisory information (signaling) to be transmitted along with VF. The two basic signaling types used in frequency-division multiplexing (FDM) systems are in-band and out-of-band signaling.

In-band signaling occurs when supervisory information is sent within the normal 300Hz to 3,400Hz circuit passband. Using a subscriber telephone circuit as an example, a 2,600Hz tone can be sent from the far-end mux channel to the central office (CO),

indicating the phone is "on-hook." When the phone is taken off-hook, the 2,600Hz signal is dropped. The CO equipment senses the absence of the tone and sends dial tone back out to the far-end.

In the case of remote base station keying, a 2,175Hz tone is frequently used. Presence of the tone activates the transmitter. The 2,175Hz signal is present for the duration of the radio broadcast to maintain the radio push-to-talk (PTT) function. The tone is within the normal VF passband of the mux circuit and is therefore considered in-band signaling. Because the tone is present during the entire broadcast, the tone must be notched out prior to the base station exciter stage, or it will be transmitted out over-the-air along with the voice.

Out-of-band signaling is achieved by moving the signaling tone frequency outside the 300Hz to 3,400Hz VF passband (but still within the 4kHz channel slot). A common example of this is the use of a 3,825Hz signaling tone.

Out-of-band signaling can also be accomplished by re-insertion of the suppressed

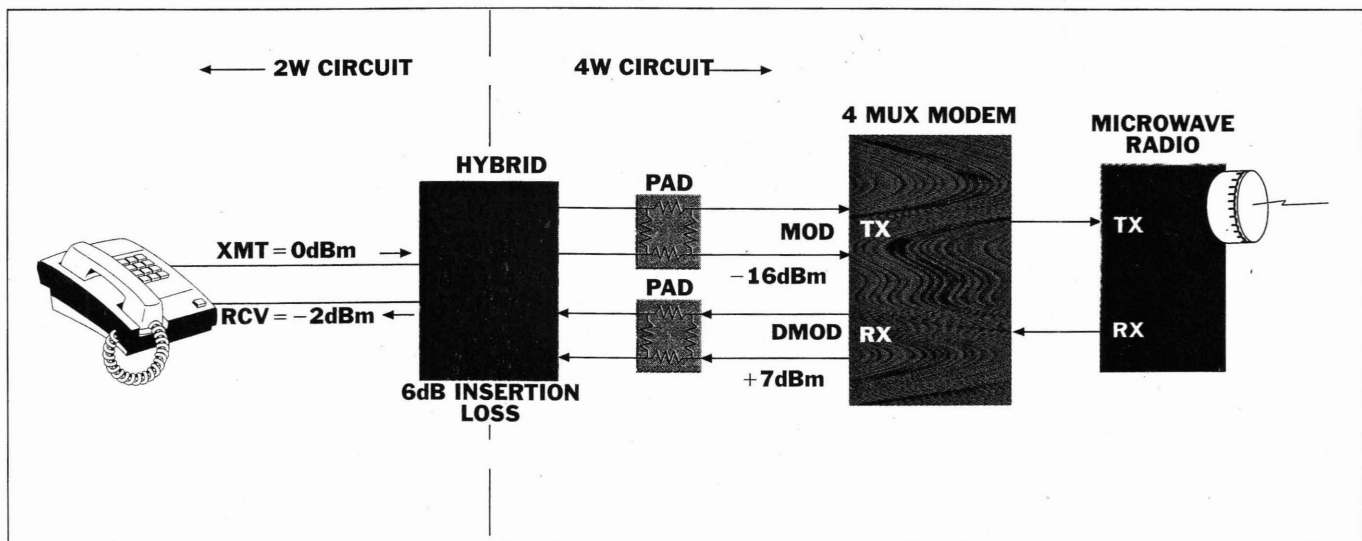


Figure 1. 2W/4W hybrid. (Signaling circuitry is not shown.)

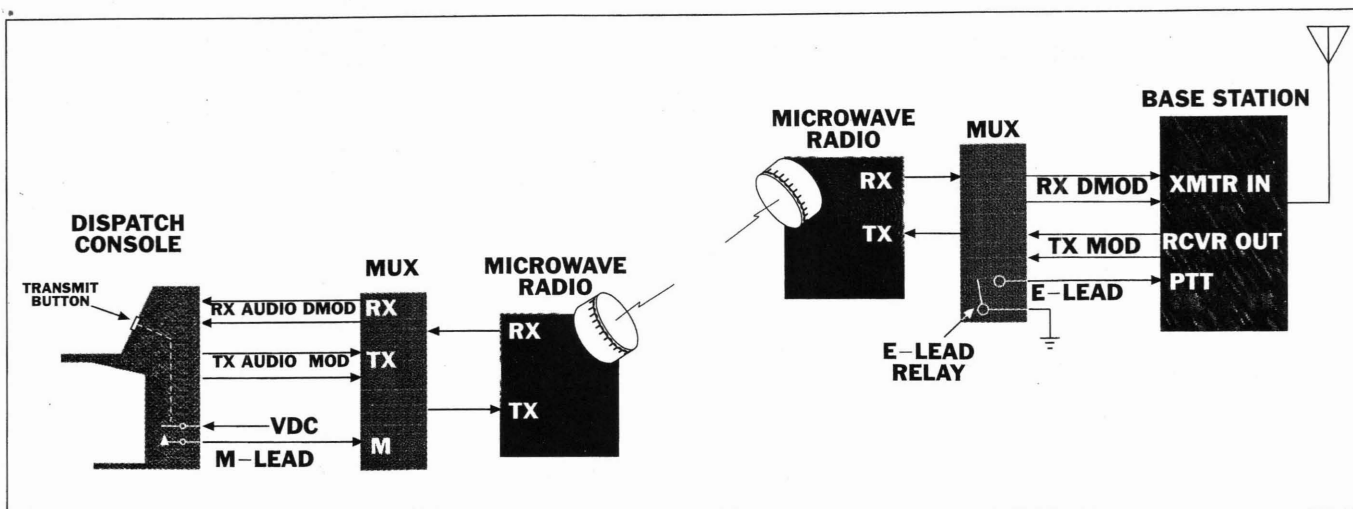


Figure 2. E-lead and M-lead signaling

carrier from the mux modem's first balanced modulator stage. Raising and dropping the level of the suppressed carrier on one end of the circuit can activate and de-activate signaling functions on the far-end. In this case, the signaling frequency will be located on the extreme edge of the allotted 4kHz channel slot.

Carrier leak is a term relating to the amount of residual suppressed carrier left in the mux modem transmit signal when the suppressed carrier level has not been raised

for signaling. Since this carrier level can be detected and used in the signaling scheme, it is important to keep it suppressed as much as possible to prevent "falsing" when signaling is not activated. Often, the mux channel modem will have an adjustment for minimizing carrier leak when signaling is not activated.

Signaling—E- & M-leads

The mux modem connections that relate to signaling are the E- & M-leads. The E-

lead connection is dedicated for receive signaling use, while the M-lead is for transmit.

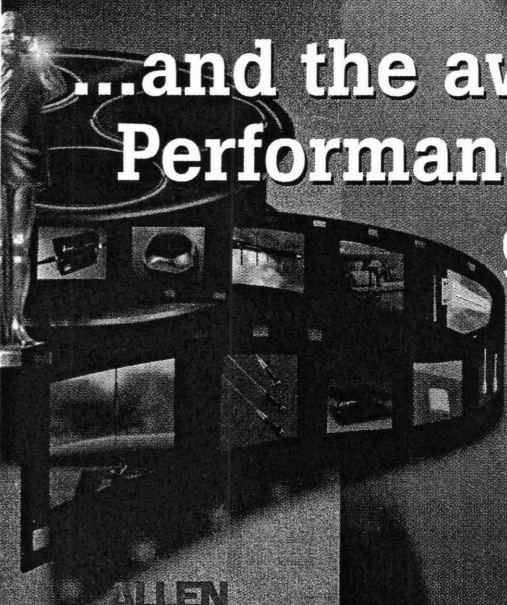
Typically, the mux modem transmits signaling to the far-end when a voltage is applied to its M-lead connection. The far-end modem detects the signaling and activates its E-lead. Usually, the mux modem is equipped with an E-lead relay that supplies external supervisory equipment with a dry set of relay contacts.

Figure 2 above offers an example of E & M signaling. A radio dispatch console

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is connected to mux so that when the console operator wishes to key up a hilltop transmitter, a button on the console is pressed. This applies a voltage to the mux channel M-lead connection, causing the modem to send M-lead signaling out to the hilltop mux via microwave. The coordinate mux modem on the hilltop detects the M-lead signal and energizes its E-lead relay. The E-lead relay contacts close and apply a ground to the radio PTT line. The hilltop base station keys up, and the dispatcher can make the broadcast.

VF transmission levels

It is important to maintain proper

signal levels within an analog multiplex system. The term *test tone level (TTL)* describes the maximum signal level that should exist at any point within an analog mux circuit to avoid creating intermodulation distortion components throughout the baseband. Test tone level is sometimes also referred to as *test level point (TLP)*.

Levels of "live traffic" can fluctuate greatly, depending on the circuit application. For this reason, they are not often a useful reference when making critical circuit level measurements and adjustments. Instead, maintenance personnel must inject an

external, constant-level test tone (typically, 1,004Hz) into the circuit. However, this is an intrusive method that disrupts normal circuit operations. As circuit gain and loss tests are performed, levels exceeding expected values are referred to as *hot*, while those which are low are referred to as either *low*, *cold* or *long*.

As shown in Figure 1, TTL for a 2W circuit in the transmit direction is 0dBm, which produces a -16dBm level at the mux MOD. If +7dBm is delivered from the mux DMOD, it should produce a receive level of -2dBm on the 2W side of the hybrid. The standard test tone level for a 4W channel modem is -16dBm into the MOD and +7dBm out of the DMOD.

In transmission systems, the term *dBm0* is frequently used to refer to the difference (in decibels) between TTL and the absolute level (in decibels relative to 1mW) being measured. For example, if a base station status tone that has an absolute level of -26dBm is applied to the MOD of a mux channel, the level is said to be "-10dBm0" at that point in the circuit. Another way to express this is to say the level at the MOD is "10dB below test tone level." Assuming proper system alignment, this status tone will come out of the far-end mux DMOD at the same -10dBm0 level. Since TTL at the DMOD is always +7dBm, the absolute level (in dBm) of the status tone can be calculated: +7dBm - 10dB = -3dBm. Even though the absolute input and outputs levels of the MOD and DMOD are different (-26dBm vs. -3dBm) they both retain the same reference to test tone level (-10dBm0). For this reason, the term *dBm0* is handy to use.

The third and final part of this series will consider VF and HF connections to analog mux systems, as well as typical test equipment found in the analog transmission systems environment. ■

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Analog multiplexing series

This is part of a three-part series on analog multiplex systems.

Part 1: "Analog Multiplex Systems: The Basics," April 1999.

Part 2: "Analog Multiplex Systems: The Basics," May 1999.

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Analog multiplex systems: The basics

Part 3 — Active analog mux systems are still in widespread use. An understanding of basic testing and troubleshooting techniques can distinguish a technician.

By Jeff Ashley

Equipped with a basic understanding of analog multiplex (mux) systems, two-way radio technicians can take the leap toward extending their duties beyond base station maintenance and into transmission systems.

This three-part series concludes with a brief discussion of mux system connections, VF and HF considerations and practical testing and troubleshooting techniques.

VF jackfield

External circuits that interface with a mux modem should do so through a VF jackfield. The jackfield allows the "end equipment" (base station, telephone, etc.) to be isolated from the mux modem for testing purposes. There should be jack positions for the MOD, DMOD, and E &

Ashley is a communications technician in Los Angeles.

M leads of each mux channel. (See Figure 1.)

The top jacks are wired to the mux channel modem and are referred to as the line side of the jackfield. These jacks allow VF level testing into the mux modem MOD or out from the DMOD. The bottom jacks are referred to as the drop side of the jackfield. These jacks allow VF level testing into and out of the drop ("customer") equipment. Without any test cords plugged in, the jackfield internally connects the line side to the drop side. If a test cord is inserted into either the line or drop jacks, the internal connection is broken and the two sides are electrically isolated.

The balanced connections to and from the VF jackfield have a circuit impedance of 600Ω. When test equipment is plugged into either the line or drop jacks, the instrument should have an internal terminating impedance of 600Ω. Because the line and drop jacks break (isolate) the two sides of the circuit, this allows the side under evaluation to be

properly terminated into the test equipment for accurate measurements. A common mistake is to have the test equipment set up in the bridging rather than terminating mode when plugged into a breaking jackfield. This will not offer the proper termination to the circuit, and it will result in level readings being +6dB hotter than what actually exists without the test equipment in the circuit.

Often, a third monitor jack is located beneath the VF line and drop jacks. Plugging test equipment into this jack will not break the circuit, but it puts the test equipment in parallel with it. In this case, the test equipment must be put in the bridging mode to avoid a double-termination to the circuit under test. A double-termination will result in readings that are -3dB lower than what actually exists when the test equipment is removed.

When measurements indicate levels are -3dB lower or +6dB hotter than expected, the test equipment terminating/bridging mode switch should be checked. The mode

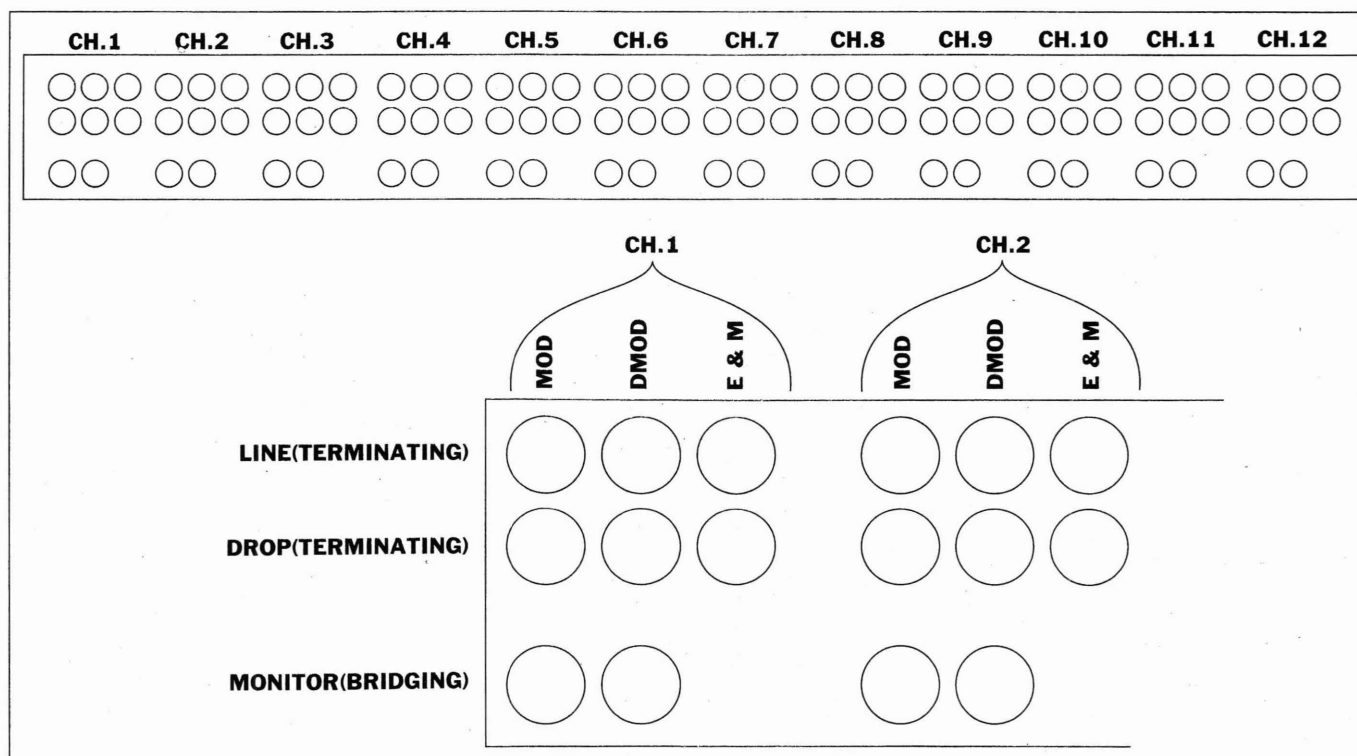


Figure 1. VF jackfield.

switch only affects the receive portion of the test equipment.

VF test equipment

Serious VF test equipment can accomplish more than just frequency and levels checks. Circuit performance parameters can be obtained from use of a transmission impairment measurement set (TIMS). Test results indicate various aspects of circuit quality. The telephone industry has established standards for many tests such as PAR (peak-to-average ratio), message circuit noise, noise-with-tone, signal-to-noise ratio, three-level impulse noise, phase jitter, envelope delay, noise-to-ground and non-linear distortion, to name a few.

HF test equipment

As the channel modems translate VF signals up to their assigned baseband frequency slots, a method is needed to measure the signal levels of the individual circuits while in the HF frequency range. A frequency selective device is required that can discriminate between the different frequencies on the baseband. Such a device is called a frequency selective voltmeter (FSVM), sometimes referred to as a selective level meter. This piece of test equipment has a variable bandpass "window" that can be tuned to any baseband frequency. It measures the power level (in dBm) of the energy within the bandpass be-

ing viewed. This allows measurement of individual multiplexed channel signals while filtering off all other baseband frequencies. The selective voltmeter can measure broadband (3kHz) or narrowband (about 200Hz bandwidth) circuit levels and typically has both flat and C-Msg weighted filters offering different response characteristics.

A spectrum analyzer is another useful tool for viewing baseband levels. There are spectrum analyzers made specifically for baseband frequencies that have their horizontal baseline calibrated to read in terms of channel, group, supergroup and mastergroup numbers. Test tone level can be calibrated on the vertical axis (signal amplitude, in dBm) so that the whole baseband can be evaluated at a glance with regard to signal levels. Hot levels can be easily identified and zoomed in on down to the channel level.

As a tool in gain/loss measurements, a bridging HF signal generator may be connected to the baseband and used to inject specific tone frequencies at specific levels. The HF frequency, when de-multiplexed, will produce a VF tone in the DMOD of the associated mux channel that occupies that portion of the baseband.

Video (HF) jackfield

The multiplex system baseband connects to a microwave radio through separate receive

and transmit coaxial cables. There should be an HF or video jackfield placed between the mux system and the microwave radio so that the two can be electrically isolated for testing and troubleshooting purposes. A monitor jack is usually provided in both the transmit and receive directions. This allows a FSVM, bridging HF signal generator or baseband spectrum analyzer to be plugged into the jackfield to gain access to the baseband without "breaking" the connection and disrupting traffic.

Potential problems

Maintaining proper signal levels is extremely important within an analog multiplex system. Whether they be externally injected test tones or "live traffic," excessive levels can drive analog mux circuitry into non-linear modes that wreak havoc on the baseband. Hot signals can not only splatter over to adjacent channel slots but produce sum and difference frequencies that affect circuits in other areas of the baseband spectrum as well. As mentioned previously, there should not be any signal on the baseband that exceeds TTL.

Another cause of baseband interference relates to poor shielding or bad solder connections in a mux/baseband assembly. Occasionally, local AM broadcast stations, which operate in the same frequency range as the baseband, can be heard on the baseband along with the mux signals. Although such interference typically affects only one or two mux channel slots, it can render those slots useless for "live traffic."

Analog mux modems also require periodic maintenance. The operating frequencies of their on-board oscillators should be held to within $\pm 5\text{Hz}$ to eliminate problems with translation error.

Conclusion

FDM systems have been in use for a long time. Although many of the latest wireless applications are employing digital technologies, active analog mux systems are abundant and employers need qualified technicians to maintain them. Many of these employers have microwave, mux and base stations at their hilltops and would much prefer having one technician "do it all." With a basic understanding of analog multiplex systems, two-way technicians will find themselves better postured in today's job market. ■

Analog multiplexing series

This is part of a three-part series on analog multiplex systems.

Part 1: "Analog Multiplex Systems: The Basics," April 1999.

Part 2: "Analog Multiplex Systems: The Basics," May 1999.

Part 3: "Analog Multiplex Systems: The Basics," June 1999.

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Circle (34) on Fast Fact Card

Tone signaling over telephone lines—A technician's primer

When planning a system that will rely on telephone circuits for tone signaling, control or access, contact your telephony provider's engineering office.

by Donald E. Koehler

If your future plans call for tone signaling or control of remote base stations, polled receivers or other communications equipment, this article is sure to be of interest to you or to your technical and maintenance staff. The technical characteristics of telephone lines, both switched and "nailed up" circuits; the kind of tests frequently used by telephony providers and some simple troubleshooting you can do using imbedded telephone tones will be covered. One word of caution: al-

Koehler teaches at the University of Alaska, Anchorage.

though the information presented here is referenced to documents from Bellcore (the telephone "standard" setter of the old Bell operating companies), always check with your local provider to see what testing you may perform and what equipment voltage levels are acceptable to their specific equipment—these vary widely in some parts of rural North America.

The telephone lines that you are most likely to use in your control scheme typically are copper wires, twisted together to form "pairs." The twist in the cable is designed to reduce crosstalk between pairs within a larger cable. If you are going to run tones and digital data over the same provided pair, let the provider know about your plans. Some types of

high-speed digital data can cause severe crosstalk. These crosstalk problems can be eliminated by line conditioning. Let's see how lines are tested and conditioned.

The cable pair(s) you use will, if unterminated, appear to your equipment as a capacitor. Even if correctly terminated, cable runs over longer distances can result in the attenuation of the higher audio or control tone frequencies. This "frequency attenuation distortion" can seriously affect your control circuit. In most cases a simple lumped inductance can balance the overall passband (frequency vs. attenuation) and provide acceptable results. If extremely long distances are used, some type of amplification is required.

This amplification is provided, in analog circuits, by a device that uses a pair of hybrid coils (see Figure 1 at the left) to "split" the signal, at least temporarily, from a two-wire circuit into a four-wire circuit.

A four-wire circuit provides for a separate transmit and receive pair to boost, then the amplified signal is placed back onto the single pair of wires by means of a second hybrid coil. (See Figure 2 below left.) This method of signal conditioning has been in use for years. The only disadvantages are that the amplifier boots both the desired signal along with any noise present on the line, singing and the possibility of echo generation caused by the hybrid coils. Checking for these "by-products" is covered later.

Connectivity to your remote site may be provided as a two-wire or four-wire switched circuit or as a permanent circuit. If the circuit is one or two pairs of wires connected between your sites without the use of switching equipment, it generally is referred to as a "nailed up" circuit. This does not preclude the telephony provider from conditioning or amplifying the line.

The "normal" passband of an unconditioned telephone line is between 300Hz and 4,000Hz—a range of frequencies that

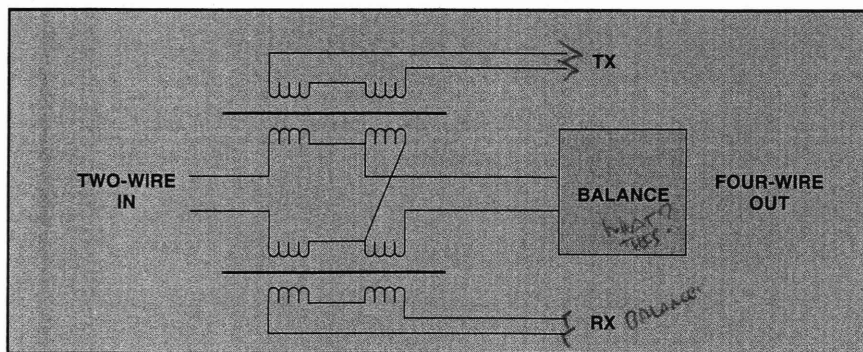


Figure 1. Using a hybrid coil to split a signal from a two-wire circuit into a four-wire circuit.

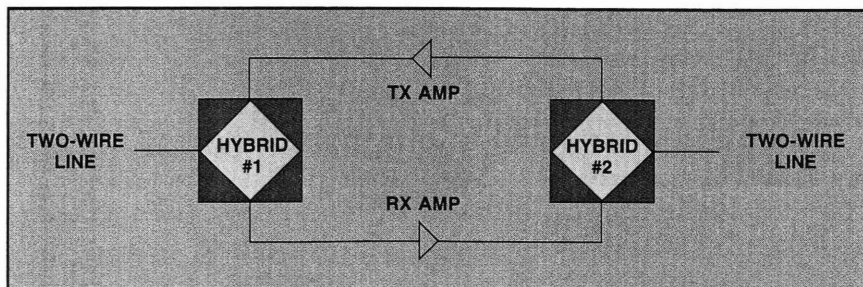


Figure 1. Boosted by a separate transmit and receive pair, the amplified signal is placed back onto a two-wire circuit using a second hybrid coil.

should offer flat response (levels) and little distortion. If your line exceeds 9 kilofeet, line conditioning equipment may be required.

Testing and maintenance of the line is generally the responsibility of the provider. Several of the tests performed by telco technicians may be of interest to you. If the circuit is switched—that is, if it runs through a modern digital switch in the local “central office,” then circuit

condition testing is mostly automated. The type of tests performed by the switch can also be performed on “nailed up” circuits with technician intervention.

The tests most likely to be of benefit to you are PAR, echo return loss and singing/singing return loss. If your needs are strictly digital data, then the bit-error rate (BER) test is the only one that is needed to test for line quality. The PAR or “peak-to-average ratio” test consists of

sending a precise set of pulses, with a known ratio of peak to average full-wave rectified voltage, into the line and measuring the resultant voltage at the far end of the circuit. Although this test will detail bandwidth, poor return loss and its attendant gain and phase distortion, the test will not define noise, jitter, IMD or other factors that can degrade control tones.

Echo on the line is the difference between the original signal and its echo, resulting from the two-to-four wire interfaces found in most central office equipment main distribution frames or transition points. Echo return loss measures loss (distortion) between 560Hz and 1,965Hz, whereas singing return loss (low) defines the distortion between 260Hz and 500Hz, and singing return loss (high) covers 2,700Hz to 3,400Hz. This is especially important if CTCSS tones must be recovered by your control equipment.

For data-only circuits, BER testing requires the use of a protocol analyzer or a

The tests most likely to be of benefit are PAR, echo return loss and singing/singing return loss. For digital data, the BER test is the only one needed to test for line quality.



Receive only

	Freq. Ranges (MHz)	N.F. (dB)	Gain (dB)	Comp. (dBm)	Device Type	Price
P30VD, P35VD, P40VD, P45VD	30-35, 35-40, 40-45, 45-50	<1.3	15	0	DGFET	\$ 44.95
P30VDG, P35VDG, P40VDG, P45VDG	30-35, 35-40, 40-45, 45-50	<0.5	26	+12	GaAsFET	\$109.95
P150VD, P160VD, P170VD	150-160, 160-170, 170-180	<1.5	15	0	DGFET	\$ 44.95
P150VDA, P160VDA, P170VDA	150-160, 160-170, 170-180	<1.1	15	0	DGFET	\$ 56.95
P150VDG, P160VDG, P170VDG	150-160, 160-170, 170-180	<0.5	24	+12	GaAsFET	\$109.95
P450VD, P460VD	450-460, 460-470	<1.8	15	-20	Bipolar	\$ 49.95
P450VDA, P460VDA	450-460, 460-470	<1.2	16	-20	Bipolar	\$ 74.95
P450VDG, P460VDG	450-460, 460-470	<0.5	16	+12	GaAsFET	\$109.95
P800VDG, P830VDG, P860VDG	800-830, 830-860, 860-890	<0.6	19	+12	GaAsFET	\$119.95

In-line (rf switched)

	Freq. Ranges (MHz)	N.F. (dB)	Gain (dB)	Comp. (dBm)	Device Type	Price
SP30VD, SP35VD, SP40VD, SP45VD	30-35, 35-40, 40-45, 45-50	<1.4	15	0	DGFET	\$ 74.95
SP30VDG, SP35VDG, SP40VDG, SP45VDG	30-35, 35-40, 40-45, 45-50	<0.55	26	+12	GaAsFET	\$139.95
SP150VD, SP160VD, SP170VD	150-160, 160-170, 170-180	<1.6	15	0	DGFET	\$ 74.95
SP150VDA, SP160VDA, SP170VDA	150-160, 160-170, 170-180	<1.2	15	0	DGFET	\$ 86.95
SP150VDG, SP160VDG, SP170VDG	150-160, 160-170, 170-180	<0.55	24	+12	GaAsFET	\$139.95
SP450VD, SP460VD	450-460, 460-470	<1.9	15	-20	Bipolar	\$ 79.95
SP450VDA, SP460VDA	450-460, 460-470	<1.3	16	-20	Bipolar	\$104.95
SP450VDG, SP460VDG	450-460, 460-470	<0.55	16	+12	GaAsFET	\$139.95

Every preamplifier is precision aligned on ARR's Hewlett Packard HP9970A/HP346A state-of-the-art noise figure meter. RX only preamplifiers are for receive applications only. In-line preamplifiers are rf switched (for use with transceivers) and handle 25 watts transmitter power. Mount in-line preamplifiers between transceiver and power amplifier for high power applications. System S/N improvement 6-14 dB typical. Other amateur, commercial and special preamplifiers available in the 1-1000 MHz range. Please include \$2 shipping in U.S. and Canada. C.O.D. orders add \$2. Air mail to foreign countries add 10%. Order your ARR RX only or in-line preamplifier today and start hearing like never before!

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load, typically 600Ω to 900Ω, or a bridging transformer, measure the voltage delivered to your site with a known or standard source providing a tone at the control end of the circuit. For switched (dial-up) circuits, dial the number for the local office "milliwatt signal." This is a tone (1,004Hz) produced at a precise level—usually 1mW (0.775Vac @ 600Ω) and terminated correctly for the equipment in use. Measure and record this (and other tone)

signal levels at both the control and line equipment (remote) end of the circuit. Leave a record of measured levels at both the local and remote site to aid in any future troubleshooting that may be required. After all, problems seldom occur during business hours.

With record in hand, you can feed a known signal to the remote site and check for excessive attenuation. In switched circuits, a call to the "milliwatt signal" from

either end of your circuit can help to pinpoint potential problems quickly. To test for possible problems with distortion requires the use of a notch filter, signal generator, ac voltmeter and some patience. If you have access to more advanced equipment, such as a sweep generator and oscilloscope, you can perform similar tests much more quickly. By setting your sweep generator to provide a signal between 300Hz and 4,000Hz, you can monitor the results at the opposite end with your oscilloscope. This approach may trigger unwanted responses from central office equipment. Central office equipment is often controlled by tones within the voice frequency (VF) passband. To avoid trouble, check with your local provider before such tests are performed. Equipment used by most telco technicians uses as few as three, and as many as 23, separate (non-sweeping or fixed) tones to perform tests for bandwidth and distortion.

When planning a system that will rely on telephone circuits for tone signaling, control or access, contact your telephony provider engineering office to:

- determine the technical characteristics of the telephone circuits available for your use.

- discuss the type of equipment and signals you plan to install and use.

- ask what support will be available for repair of potential problems.

- find out what kind of testing you will be permitted to perform and how to best express this information to the telephony provider's technical staff.

- confirm the costs of any technical assistance required and provided, and the limits of responsibility.

Document these discussions, and provide a copy to your provider. If a problem does arise, records are always better than a memory—and much safer as well.

Acknowledgement

The Bellcore "Redbook" of engineering standards was used as a reference in the preparation of this article.

Recommended reading:

Clarke, Martin P., *Networks and Telecommunication—Design and Operation*, John Wiley and Sons, ISBN 0-471-92799-6.

Visit the Bellcore web site, (www.bellcore.com), and the "Telephony 101" document on the Northern Telecom site (www.nortel.com) to learn more about the public switched telephone network (PSTN).

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Circle (16) on Fast Fact Card

Many out of one: Multicoupler basics

By Harold Kinley

What do you do when your "boxes" outnumber your available "sticks"? In this business, it is often necessary to feed more than one receiver from a single antenna connection. This is generally referred to as *multicoupling*. Techniques to accomplish this vary widely, depending on the available signal level and other factors such as required isolation and good engineering practice. Methods of multicoupling range from the simple to the complex, so let's start with the simplest arrangement.

A direct connection does not maintain the 50Ω system impedance and does not provide any isolation between the receiver ports (tee connector). This arrangement is only used when lack of isolation is not a problem and the signal level is fairly high (not near the receiver threshold sensitivity).

More than two receivers can be combined by using more tee connectors. For example, two tee connectors can be used to feed a single antenna to three receivers as shown in Figure 2 on page 32. The number

ports. The loss will be greater for dividers with more output ports. The loss between the input and any output port can be calculated as

$$L = 20 \log \left[\frac{1}{N} \right]$$

where

L = loss in decibels

N = number of output ports

The system impedance will be maintained, and some small amount of isolation will be between receivers. However, the resistive-divider network is inefficient and usually an undesirable way to provide multicoupling.

Transmission-line transformer

To maintain a constant 50Ω system impedance, a quarterwavelength transmission line of the proper characteristic impedance can be used as an impedance transformer. For example, to feed the input of two receivers from a single antenna, using a simple tee connector as a splitter will reduce the impedance at the splitter input to about 25Ω. This is because the two 50Ω receiver inputs are connected in parallel and thus produce a 25Ω impedance.

If the 50Ω impedances of the receiver inputs were to be transformed to 100Ω and then combined at the tee connector, the resultant impedance would be 50Ω. This would maintain the integrity of the 50Ω system impedance. You can closely approximate this by using a quarterwavelength of 75Ω transmission line. To transform the 50Ω receiver-input impedance to 100Ω would require a quarterwavelength of transmission line with a characteristic impedance of 70.7Ω. This is calculated as:

$$Z_0 = \sqrt{50Z}$$

where

Z_0 = the characteristic impedance of the transmission line

Z = the impedance to be transformed. This formula also can be rearranged as:

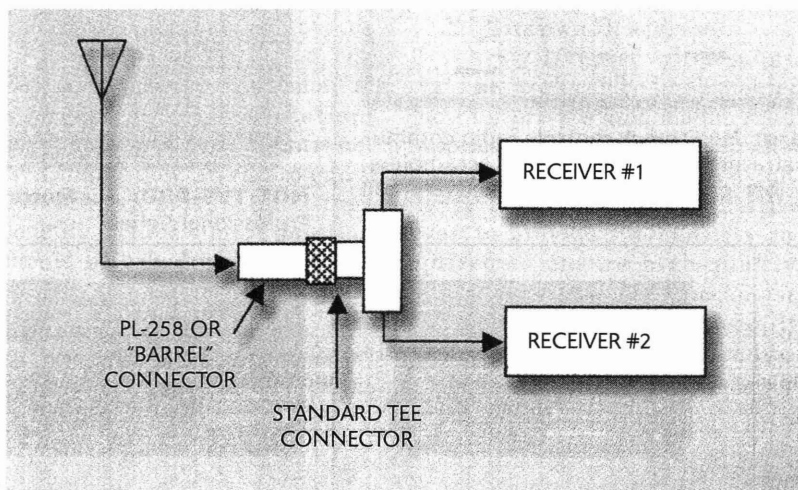


Figure 1. The simplest way to feed two receivers from a single antenna is to use the "tee" connector arrangement shown here. It should only be used when isolation between receivers is not necessary and the signal level is adequate for division between the receivers.

"Tee" connector method

In the basic setup shown in Figure 1 above, we need to feed the inputs of two receivers from a single antenna. A simple "tee" connector is used to split the antenna line for each receiver input. This type of

of tee connectors required to provide N outputs is equal to $N - 1$.

Resistive network method

Another simple method used to provide multiple outputs from a single source is the resistive method. The size of the resistors, such as in Figure 3 on page 32, is calculated as:

$$R = Z_0(N - 1)$$

where

Z_0 = system impedance (50Ω)

N = the number of outputs

The loss between the input port and either output port will be 6dB for the divider with two output

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$$R = Z_0(N-1)$$

$$R = 50(2-1)$$

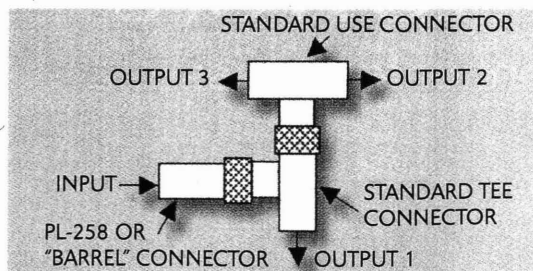


Figure 2. An input signal split three ways with two tee connectors. The number of tee connectors required is $N-1$, where N is the number of desired outputs.

$$Z = \frac{Z_o^2}{50}$$

This will provide the resultant impedance when using a quarter-wavelength section of transmission line of characteristic impedance, Z_o , on a device with a 50Ω impedance. Transmission lines with a characteristic impedance of 70.7Ω are not readily available, but 75Ω transmission lines are readily

available. So, substituting 75 into the formula we have:

$$Z = \frac{75^2}{50} = 112.5\Omega$$

This means that connecting a quarter-wavelength section of 75Ω transmission line to an impedance of 50Ω will yield an impedance of 112.5Ω at the input, as shown in Figure 4 on page 34. This impedance of 112.5Ω is purely resistive, and it is properly written as $112.5 \pm j0\Omega$.

Now, to combine two 50Ω receiver inputs, we could use two quarter-wavelength lines, as shown in Figure 5 on page 34, to produce an impedance of 56.25Ω at the input to the tee connector.

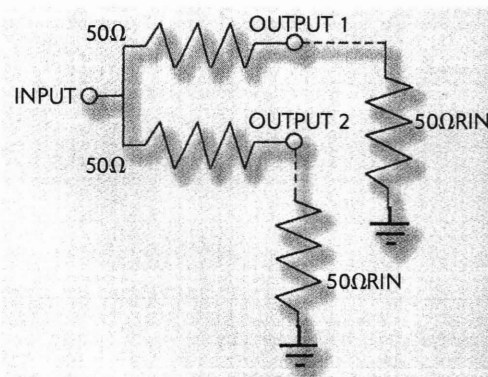


Figure 3. Two 50Ω resistors form a splitter network to feed two receiver inputs. The number of resistors needed to make such a resistive splitter equals the number of outputs needed. The value of the resistor is $Z_o(N-1)$, where Z_o is the system impedance (50Ω for our purposes) and N is the number of outputs needed. For two outputs, we need 50Ω resistors in the divider network. The two 50Ω resistors shown connected to the outputs of the divider network represent the input impedance of the receivers.

This would produce a VSWR of $1.125:1$ —a good impedance match.

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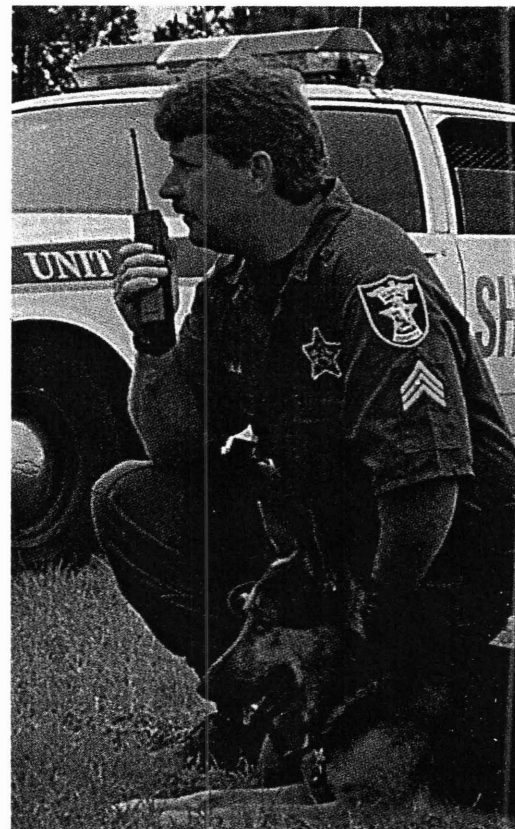
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Technically Speaking

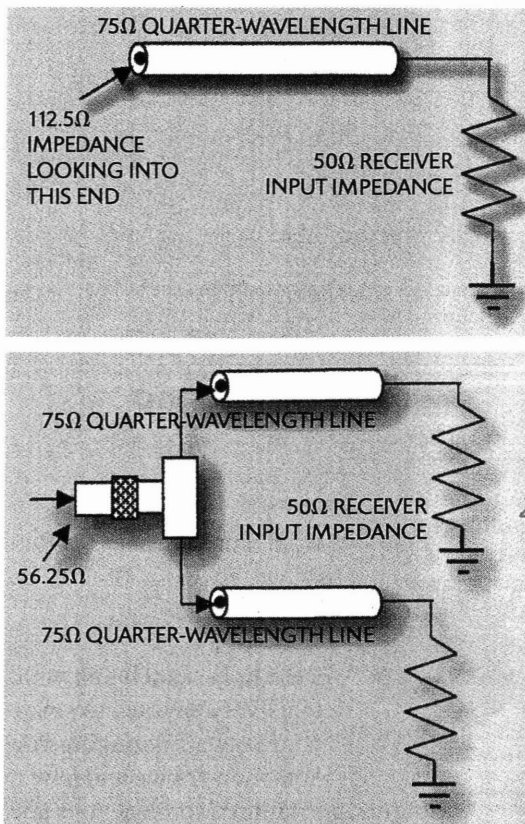


Figure 4 (top). A quarterwavelength of 75Ω transmission line transforms a 50Ω impedance into an impedance of $112.5 \pm 70\Omega$ at the other end of the transmission line. Figure 5 (bottom). Two quarter-wavelength 75Ω transmission lines provide an impedance match to a 50Ω system while splitting the input to two 50Ω receivers. The impedance 'looking into' the input side is 56.25Ω. It provides a close match to the 50Ω system impedance.

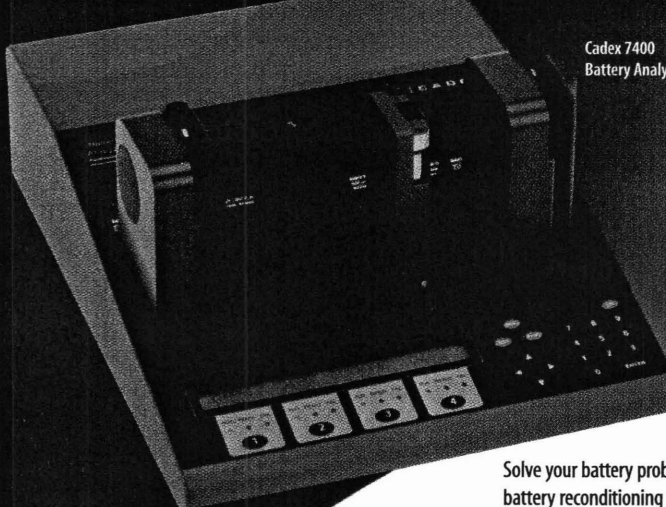
When cutting the transmission line to the proper quarter-wavelength, it is necessary to account for the velocity factor of the transmission line. Remember, too, that odd multiples of quarter-wavelength may be used, for practical purposes.

To divide the input into three outputs, quarterwavelength sections of RG-62 coax could be used. The characteristic impedance of RG-62 is 93Ω. Terminating it with the 50Ω receiver input will transform the impedance to 173Ω at the other end. Three of these connected

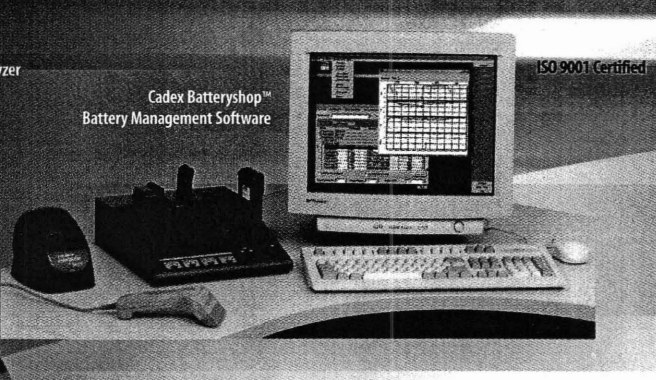
in parallel would produce an impedance of $173 \div 3 = 57.6\Omega$ for a VSWR of $56.6 \div 50 = 1.132:1$ at the input to the divider. This is close enough for practical purposes. Similarly, a 1:4 divider could be fashioned using four quarter-wavelengths of RG-62. Because $173 \div 4 = 43.25\Omega$, this will produce a VSWR of $50 \div 43.25 = 1.156:1$ at the input to the divider.

A commercially produced receiver multicoupler is the recommended way to go. Such multicouplers are usually designed as rack-mount units with a built-in preamplifier and hybrid power splitter, along with the necessary power supply. Dividing the input signal among several receivers will necessarily reduce the amount of signal level available at each output port.

For example, the *theoretical* insertion loss of a two-port divider is 3.0dB, for a three-port divider, 4.8dB, and for a four-port divider,



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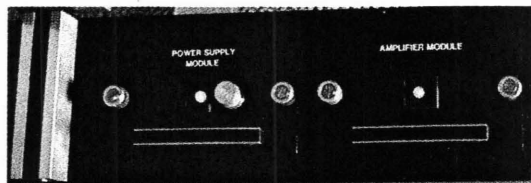
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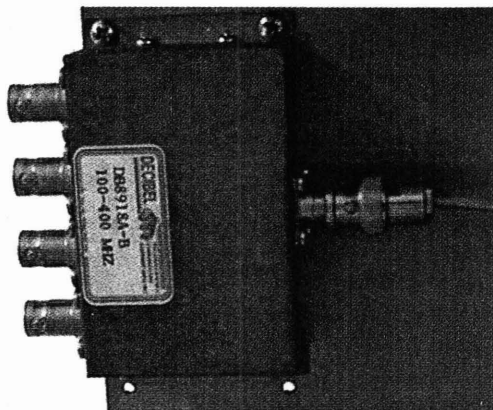
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Technically Speaking



A commercial-grade multicoupler unit.



A commercial multicoupler's signal divider is usually mounted on the back of the chassis.

6dB. In practice, the insertion loss will be greater than that, but it can never be *less* than that. Therefore, if you see an ad specifying inser-

tion loss *less than the theoretical minimum*, you can usually take that to mean that the insertion loss is that amount *plus* the theoretical minimum.

In commercial multicouplers, a low-noise preamplifier usually precedes the power divider, as shown in Figure 6 at the right. This will provide the best noise figure for the system. A high degree of isolation is also provided between the output ports for receiver-receiver isolation. When purchasing a commercial multicoupler, look closely at specifications such as: frequency range and bandwidth, receiver-receiver isolation, third-order intercept point, VSWR, system gain, noise figure, power requirements and preselector.

A commercial multicoupler unit includes a power supply and preamplifier, which are modular and can be easily removed for servicing. A signal divider is usually located on the rear of the chassis. A low-noise, high-gain preamplifier is used to overcome the losses of the

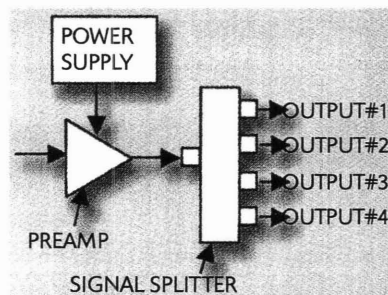


Figure 6. A typical receiver multicoupler provides a low-noise preamplifier preceding the signal divider. The amplifier should also have a high third-order intercept point for good IM rejection. The divider should provide a high degree of isolation between output ports. Any unused output port should be terminated in a 50Ω impedance.

divider while maintaining a low noise figure.

Homemade signal dividers might be useful in a given situation, such as an emergency. However, to maintain a system with optimum performance, use a commercial-grade multicoupler unit.

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Technical techniques: A primer for transmission lines

Part I—An understanding of transmission lines and tips on using them as transformers and filters can help techs properly configure feedlines and even solve some problems.

By Patrick E. Buller

Transmission lines carry RF from point to point within radio equipment, between pieces of equipment and from equipment to antennas. They also can be configured as filters and impedance transformers. In simple

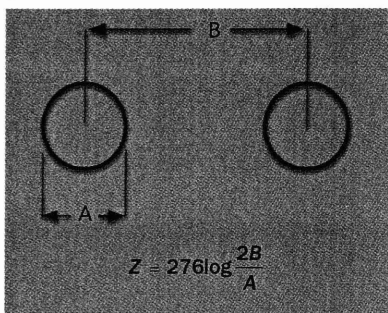


Figure 1. Balanced two-wire lines: formula for determining impedance of a two-wire transmission line. (A and B are the same units.)

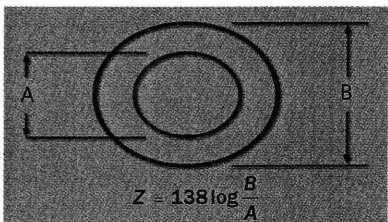


Figure 2. Characteristic impedance of an air-insulated coaxial line can be found by the above formula. (B and A are the same units.)

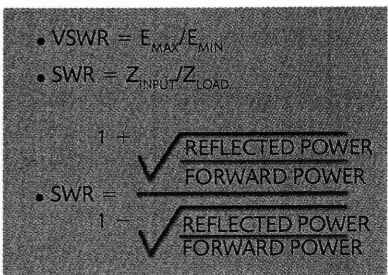


Figure 3. Standing wave ratio.

terms, a transmission line connects the source of energy to the load.

Transmission lines are either *balanced* or *unbalanced*. A two-wire transmission line becomes balanced when each conductor has equal impedance relative to ground or to its surroundings. An unbalanced transmission line has one conductor grounded, or exposed to the outside elements in greater magnitudes than the other conductor.

The impedance value is a function of conductor diameter and spacing between the two conductors. Impedance can be lowered by increasing the diameters, decreasing the spacing between the two conductors or both (see Figure 1). The most common balanced line impedances of today are 300Ω and 450Ω twin lead and 124Ω twinax (balanced coax cable). Seventy-five-ohm cable was common prior to the 1950s because most wire antennas were halfwave center-fed dipoles with impedances near 75Ω.

In an emergency, common 16 gauge lamp (zip) cord would suffice as a dipole transmission line. However, the sun deteriorates this type of cable when used for years at a time. Open-wire transmission line is more common for ham radio stations, not commercial or public safety facilities.

Coax cable

Co means "same or identical," and axial-axis means a "center point or point of common rotation." Coaxial cable is an unbalanced transmission line. Figure 2 shows the relationship of two conductors and the formula for determining

the impedance of a coaxial line.

The center conductor must be kept evenly spaced inside the outer conductor, usually by a *separator*, or *media*. Common media include ceramic insulators (beads), spiral dielectric, polyethylene, fluorinated ethylene propylene-FEP (known as foam) Teflon and air or inert gas with separator beads.

Velocity factor

The media surrounding the center conductor slows the travel of RF energy. How much depends on the material used. The ratio of the time it takes for RF to travel through free space compared to the time it takes to travel through cable is called the velocity factor *vp*. The lower the value of *vp*, the longer it takes RF to reach the end of the cable. Common *vp* values are 0.66 for standard cable, 0.78 for foam, 0.82 for airfoam and 0.98–0.99 for air or inert gas.

All cables cut to a wavelength will be physically shorter than a free-space wavelength. For example, a halfwave at 150MHz is 39.36 inches. Using RG8/U cable, (*vp*–0.66), the physical halfwave cable length would be 39.36 × 0.66 = 25.98 inches.

Velocity factors become important when coax cables are used to feed phased arrays and to interconnect cavities and duplexers. They

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are also important when a time domain reflectometer is used to find cable faults.

Impedance

A transmission line impedance can also be determined by $(L/C)^{1/2}$ or by Ohm's law, and would be found by measuring the voltage and current at the sending end of a line of infinite length. In practical terms, for cables shorter than infinity, an adjustable resistance can be added at the end and adjusted to a value that provides the same input voltage and current relationship of the infinite length. This value when measured, will equal the characteristic impedance of the line.

A second method for determining impedance is to terminate any length of cable in its characteristic impedance, resulting in no reflected energy. That way, maximum transfer of energy takes place. Any load connected to a transmission line with an impedance different from the line's characteristic impedance results in a standing wave pattern.

The input of cable measuring hundreds of wavelengths would have virtually no reflected energy and would appear to have an impedance input equal to its characteristic impedance. An example is 500 feet of RG58/U fed with 450MHz of energy. It would show an impedance of 50Ω and minimal reflected power because almost all of the energy would dissipate as it travels the length of the line. It would be an inexpensive dummy load.

The cable TV industry uses this concept in multiple distribution points where taps on the coax cable appear along the length of the cable. An amplifier supplying several volts of signal at the head end is sufficient to overcome the losses, and the loss of the cable terminates the amplifier.

VSWR

Voltage standing wave ratio was the first method for transmission line measurement. If the load

absorbs all of the energy sent down the transmission line, no energy returns.

To measure this effect, a detector (often a voltage probe) is moved along the line, maintaining uniform distance from the line, noting the voltage change, if any, as the probe moves along and parallel to the transmission line. The change V_{max}/V_{min} equals the VSWR. The purpose of this test is to determine how closely the load matches the line. The distance between each voltage minimum equals the halfwave length of the transmission line frequency, not the applied frequency. This measured distance, compared to the applied frequency (wavelength), yields the velocity of propagation, vp . The movable probe method is used today mostly as a laboratory method and is most suitable when using the Smith Chart.

Determining SWR by reflected power

Today's methods for measuring forward power and reverse power at a point on the transmission line include:

- applying the formula shown in Figure 3.
- measuring forward and reverse power in dB and comparing the ratio of dBs, i.e. 25dB return loss.
- measuring power with a reflectometer and adding to the power reading the coupling loss of the device.

Reflected power is 180° out of phase from the forward wave as shown in Figure 4. To demonstrate this phenomenon, attach a rope or line to a fixed point, like a wall or doorknob, and with your wrist thrust a forward-looking wave. Note that immediately thereafter, a reflected wave will come back to you in the reverse mode. This demonstrates the principle by which directional couplers operate.

Figure 5 is a diagram of a directional coupler that consists of two short transmission lines in close proximity. Line L2 is a transmission line that has its own charac-

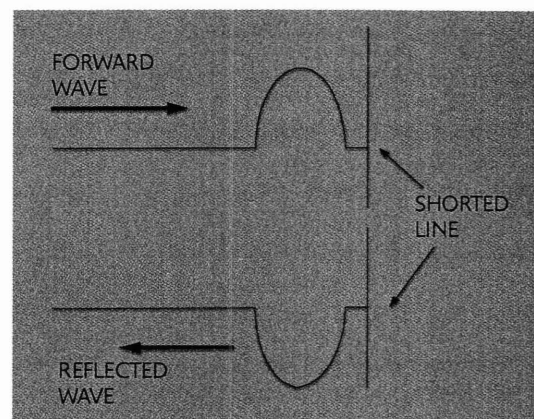


Figure 4. Reflected wave is 180° out of phase from the forward wave. It may lead or lag, depending on type of reactance load.

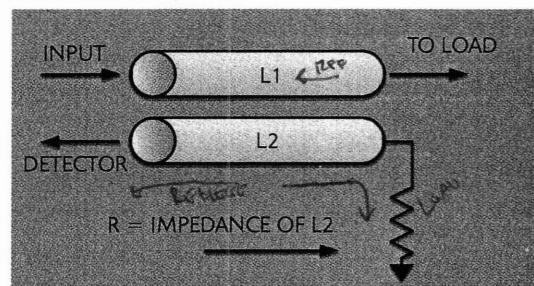


Figure 5. Directional coupler: This arrangement detects reflected power. Forward power can be measured by reversing power flow in L1.

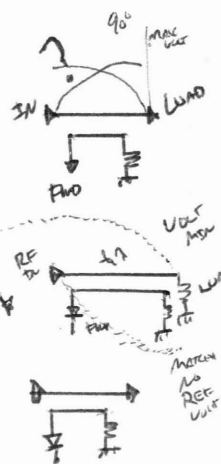
teristic impedance terminated by a resistor R . Power traveling in L1 from left to right induces energy in L2, but this energy is dissipated in resistor R .

If the load is not equal to the impedance of line L1, energy will be reflected back on L1 and in turn coupled to L2 and on to the detector. Reflected power in this example is measured.

If the input power and load were connected in reverse, the detector would then measure forward power.

If L2 were physically reversed, keeping the input power on the left, forward power would then be measured.

This example explains how popular wattmeters operate. L1 is the center conductor of the transmission line of the wattmeter. The rotatable slug is made up of L2, resistor R and a diode that connects to the meter calibrated in watts. Rotating the slug samples power in either the forward or reverse direction. Note



Transmission Line Basics

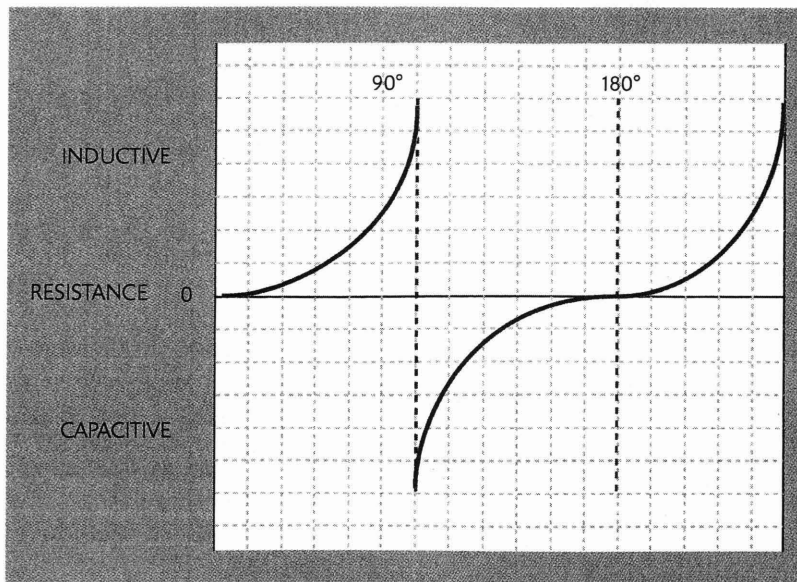


Figure 6. Graph showing shorted transmission line.

that the direction arrow on the slug corresponds to the detector side of L2, in Figure 5.

Directional couplers are useful in checking site noise or repeater de-sense. The benefit is that the reference signal generator is sent to the receiver, not to the antenna. Directional couplers are available from megahertz to gigahertz, the

latter in the form of rigid waveguide components.

Transmission lines as stubs

A transmission line, when added to another transmission line in either series or parallel, can provide impedance matching. A transmission line that has a short circuit on one end will exhibit reactive

components depending on its electrical length as shown in Figure 6. A quarterwave line shorted at one end has a high impedance at the opposite end—meaning it can be connected in parallel on a transmission line of the same impedance without affecting the line. If a halfwave shorted line were connected, the effect would be the same as placing a short on the transmission line at that point.

Another way of looking at this concept is that a quarterwave stub shorts out any second (or even) harmonic. All odd harmonics will be seen as high impedance.

Referring to Figure 6, the following can be stated: A short on one end of a quarterwave is open at the other end. A short on the end of a halfwave is a short at the other end.

The same result is observed when the cable is lengthened in any multiple of half-wavelengths.

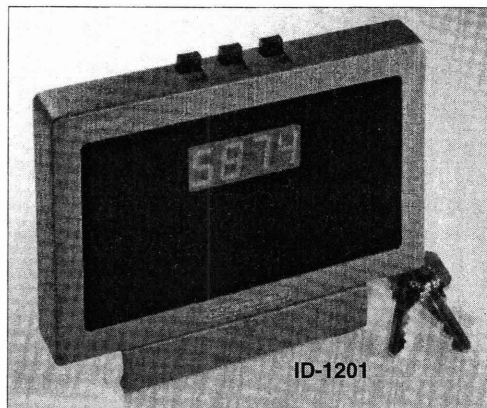
The phase of the waveform shifts by 90° at a quarterwave, 180° at a halfwave and 360° at one wavelength of transmission line. Transmission line stubs can therefore be either open circuit, short circuit or terminated—all too often, poorly terminated.

Transmission lines as helpers

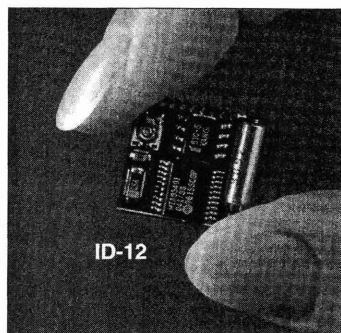
A length of open-end coax connected as shown in Figure 7A adds reactance according to its electrical length. Coax cable can be changed artificially by adding capacitive reactance to the open end of a stub to achieve the exact reactance desired as shown in Figure 7B. This short length of cable can be "tuned" by a variable capacitor connected between the center and outer conductor on the "open end" that electrically changes the cable length. The net effect is a series, low-impedance circuit connected at that point.

This is also a handy device for tuning a transmission line to measure what effect a cable would have when cut to the exact length. The capacity-tuned line is often used for tuning the coupling link of a cavity. It usually consists of a rigid

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outer and center conductor with a movable dielectric that has the same effect as a variable capacitor.

A tuned cable connected by a "tee" can reduce interference when there is at least a 2:1 frequency difference between the desired signal and the interfering one. The bandwidth is a function of the diameter of the cable.

Transmission lines as transformers

The quarterwave transformer is useful for changing impedances. It can take the form of either a balanced line or a coax transmission line. Use the formula $Z_{line} = (Z_a \times Z_b)^{1/2}$, where Z_a is the impedance desired and Z_b is the impedance to be matched. Combining two 50Ω impedances yields 25Ω . $(50 \times 25)^{1/2} = 35.5\Omega$ cable. This result can be achieved with two parallel 72Ω cables or with 35Ω cable. In either case, this 35.5Ω quarter-wave electrical length of cable connected to the parallel 50Ω devices transforms 25Ω to 50Ω.

Another way of solving the same problem is to change each of the 50Ω devices to 100Ω so that when they are connected in parallel, the net is a 50Ω point. The quarterwave cable would then have the impedance of $(50 \times 100)^{1/2} = 71\Omega$. Any of these methods can be used. They are found mostly in antenna-combining units where two or more antenna elements are combined to one 50Ω common feedpoint.

Another method of impedance matching is to make the quarter-wave transformer out of rigid tubing. The trick is to find the right diameter for the center conductor. Using the previous example, here's how to find the right size for the hardware. Assume a quarterwave-long 35Ω transformer is desired, and we have a tube with an outer diameter of 1 inch, inside 0.78 inch. Use the formula of Figure 2 with Z_o of $35 = 138 \log B/A$ where B is 0.78 inch and A is found to be 0.435 inch. A $7/16$ inch rod would serve the purpose.

The complete transformer will be a $3/4$ inch pipe with a $7/16$ inch

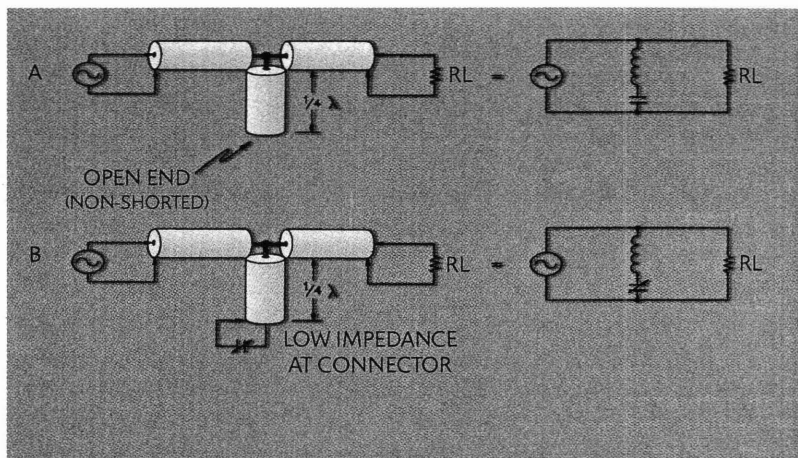


Figure 7a. Open-end stub: Not a complete open circuit, it has some radiation. 7b. Increasing capacity will lower frequency of low impedance. This method has a lower Q.

diameter center rod with a coax fitting on one end of this coax line, and the opposite end will have two coax connectors in parallel to this quarterwave rigid pipe. The combination forms a matching transformer with high-power capabilities. These devices are listed in catalogs as power splitters or hybrid combiners. Now you know the secret to making your own, inexpensively.

Balanced transformers

A further modification of the quarterwave transformer gives it a balanced output. If another larger diameter pipe were placed over this 1 inch pipe, say a $1\frac{1}{2}$ inch pipe (making the combination "tri-axial"), and if it were grounded or connected to the feedpoint, the result would be a balanced 35Ω output because of the decoupling effect of the $1\frac{1}{2}$ inch pipe as shown



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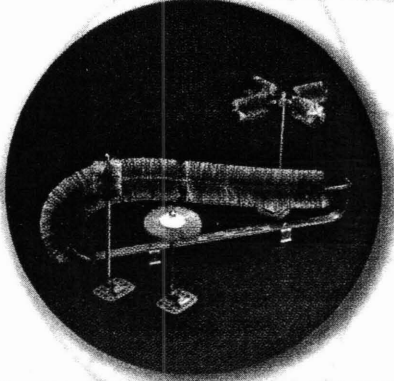
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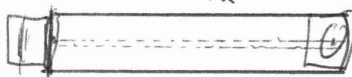
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Transmission Line Basics

in Figure 8A.

Coax baluns (BALANCED to UNBALANCED) offer another method of obtaining balanced lines as shown in Figure 8B and 8C. Figure 8B is a 1:4 impedance step-up because the current divides at the

junction point, leaving half of the total current feeding each line of the balanced port. In this example, the output impedance would be 200Ω. Figure 8C has no current division, therefore offers a 1:1 impedance transformation. The

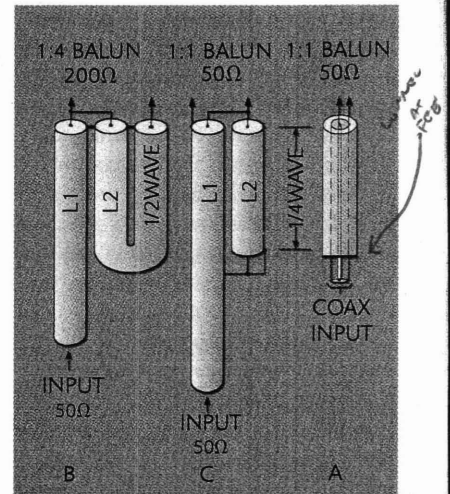


Figure 8. Coax balun. BALANCED to UNBALANCED. $Z_L = L1 = L2$, all cases.

additional coax cable alongside converts the common ground point to high impedance because of the quarterwave transformation from a short to an open. These kinds of balanced transmission line transformers are effective tools in matching broadside arrays fed with coax cable.

With a basic understanding of transmission lines and some helpful tricks for using them as transformers and filters, you can configure feedlines properly and solve some problems along the way. ■

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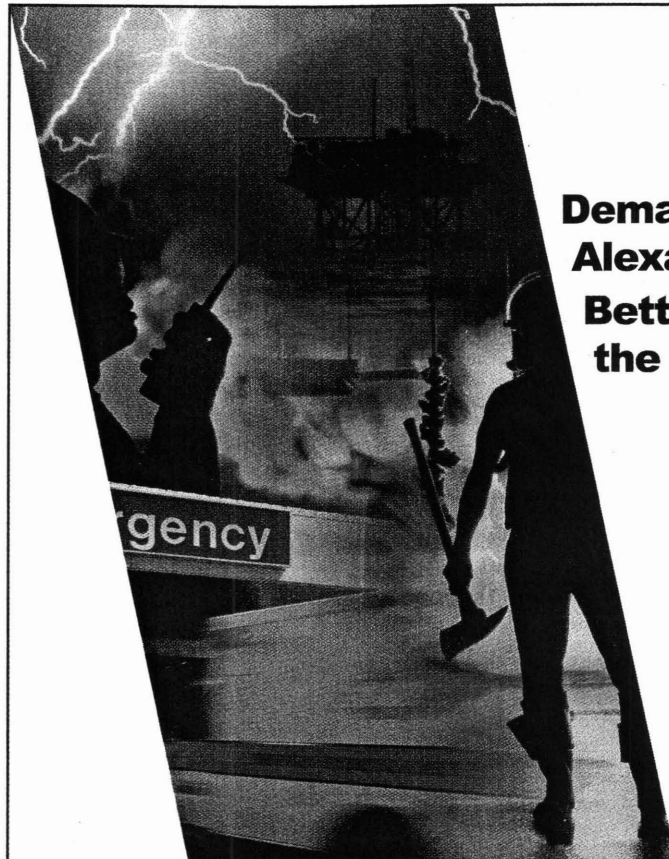


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Technical techniques: A primer for transmission lines

Part II—An understanding of transmission lines and tips on using them as transformers and filters can help techs properly configure feedlines and even solve some problems.

By Patrick E. Buller

One of the most fascinating types of transmission lines is the *waveguide*. Radio waves—if confined within a hollow tube of conducting material and if the dimensions of the tube are

Rigid waveguide

Rectangular is the most common waveguide form in the radar industry. Elliptical is used for terrestrial microwave. Circular was most popular for the long-haul, high density, point-to-point microwave service such as AT&T. Radio energy travels within this confined area by reflecting off the sidewalls of the guide. The reflections occur so that the electric and magnetic fields of the waves set up a definite pattern within the guide. These patterns are called *modes*.

In addition, every mode has a *cutoff frequency*. This is the lowest frequency that will propagate down the waveguide. The lowest frequency that can travel through the waveguide is called the *dominant mode*. A dominant mode is the frequency that has a wavelength equal to twice the width of the guide. This mode has the simplest field pattern and is not as susceptible to impedance mismatches and reflections as the more complex modes. Usually, a rectangular guide has a width greater than half a wavelength but no more than one wavelength, and the height is half this width.

A rectangular waveguide can be thought of as a special case of a two-wire transmission line supported by quarterwave shorted stubs as shown in Figure 1. The voltage is a maximum at the center point corresponding to the imaginary two wires and a voltage minimum at the shorted ends.

Figure 2 shows the transverse electrical TE 10 mode. It's characterized by the fact that the

electric vector (E vector) is always perpendicular to the direction of propagation. Note that the electric "E" field exists within the waveguide and is maximum in the center and zero at the ends. From this we can see that the "E" field is the voltage field.

Energy is inserted or removed by a simple probe spaced a quarter-wavelength from a shorted end of waveguide and at the center of the guide as shown in Figure 3. The probe is considered a quarterwave antenna with ground plane because the entire waveguide is made of conductive material, usually oxygen-free high conductivity copper. Remember, to make an antenna broadband, increase its diameter or add capacity. Most coax-to-waveguide adaptors have the quarterwave antenna captured in a polystyrene tube that not only fixes the probe, but also adds a sufficient amount of capacity making it a broadband antenna.

A contributing factor of efficient wave propagation within a hollow waveguide is not having any abrupt change in the physical dimensions. In other words, dents or foreign material within the waveguide disturb the E and H field causing reflections that result in a standing wave pattern. Any system using a waveguide is most

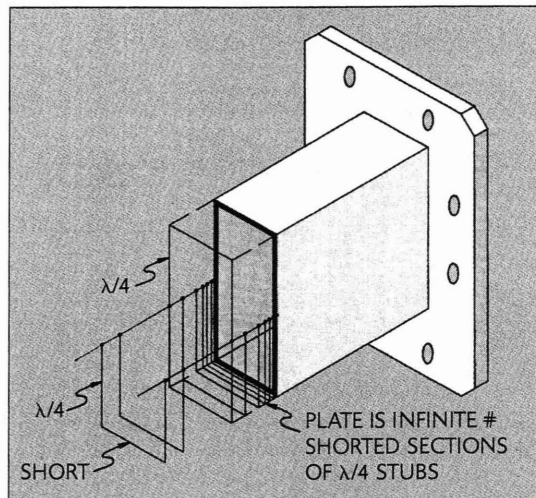


Figure 1. Waveguide—derivation from two-wire balanced line.

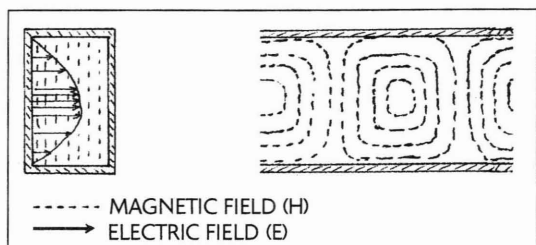


Figure 2. Field configuration for a TE 10 wave.

appropriate for the frequency of operation—will travel the length of the tube with little loss. Such a tube is a waveguide. It comes in three physical forms: rectangular, elliptical and circular.

Buller is a special projects engineer for Tacoma Power, Tacoma, WA. For many years, he served as an electronics design engineer for the Washington State Patrol. He is a member of IEEE, NARTE, APCO and ARRL, and he is a Fellow of the Radio Club of America. His email address is W7rqf@msn.com.

vulnerable to damage here. A sign often seen on radar installations is "Warning-Waveguide-Do not chip or dent." This warning is valid whether the waveguide is circular, rectangular or elliptical.

An exception to this rule is when probes are installed for tuning purposes. A machine screw installed in the center of the guide will either add capacity or inductance, depending on its placement within the VSWR pattern. Once the "E" field is disturbed, it remains disturbed all along the line until it reaches another point of opposite disturbance where it can be canceled. Tuning screws are frequency dependent and will have different effects depending on where they're placed, not only along the line but how deeply they penetrate. Small dents appropriately placed on a rigid waveguide

can add opposite reactive components that cancel those within the waveguide. The experienced person can tune a waveguide by carefully applying pressure at precise points along a waveguide while observing the return loss. Waveguide components such as the E and H bends often have a decal near distortions stating "dent tuned," meaning, don't call us because of apparent damage.

Adding several tuning screws close together ensures tuning ability for either kind of reactance. Tunable elliptical waveguide connectors have tuning screws where the transition is from elliptical to rectangular. These tuning screws take care of the connector discontinuities and do little for transmission line discontinuities many wavelengths away.

By observing the H field of

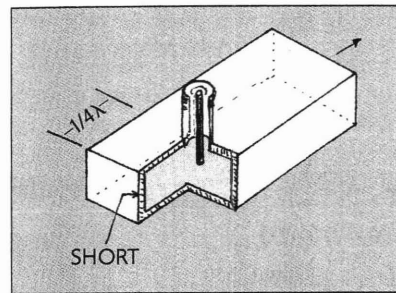


Figure 3. Cut-away showing probe coupling to waveguide.

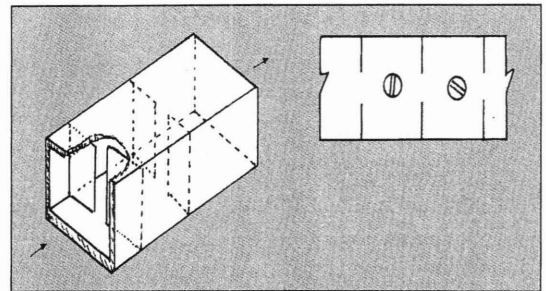


Figure 4. Resonant cavity from waveguide.

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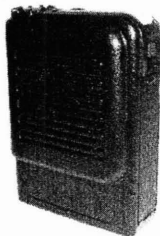
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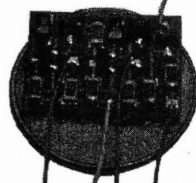
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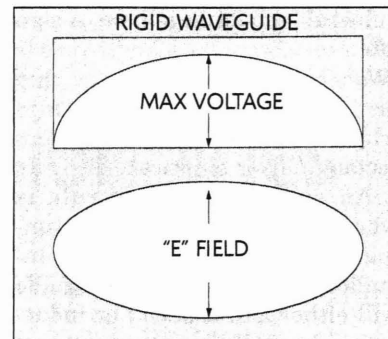
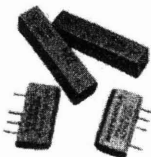


Figure 5. Elliptical waveguide follows voltage gradient.

Figure 2, we see that this field has a definite pattern. If the interior of the waveguide had walls spaced identical to the field pattern, a cavity could be made out of this guide. Indeed that is what is done. Walls placed within the guide (as shown in Figure 4) add inductance, and a tuning screw adds capacity that will tune this area to a specific (waveguide) frequency. Energy travels from one compartment to the other (aperture coupling), but only at a frequency that is resonant to the $E \times H$ field.

This method of tuning a waveguide is usually used for "front ends" of a microwave transceiver. Its bandwidth is a function of how many "rooms" and how large the door is between rooms.

As in coax transmission lines, waveguide transmission line cavities described above have low "Q" values, (less than 50), meaning it's a broadband device. High Q values cannot be obtained with waveguide dimensions and are constructed as large cylinders connected to the waveguide. High Q values are required for reference oscillators or AFC cavities.

Elliptical waveguide

Elliptical waveguide is an efficient modification of rigid or rectangular waveguide. Its physical shape tends to follow the "E" field shown in Figure 5. The advantage of the elliptical waveguide is manufacturing convenience. It's composed of a strip of copper that is corrugated, folded over, welded and jacketed. It's available in continu-

ous lengths up to shipping limits of around 3,000 feet. The only joints are the connectors on each end. Elliptical waveguides can be bent in either the E or H plane without hardship or damage. Rigid, on the other hand, was limited to 20-foot lengths requiring choke and cover flanges at each 20-foot distance. Prefabricated 90° bends were needed (two more flanges) to change directions.

All hollow waveguides need air under pressure (3psi-5psi) to prevent moisture migration to the inside. Any leak allows air to escape and prevents moisture from getting in. For high-power installations, 15psi of pressure or more raises the voltage level that the waveguide will handle without internal arc-over.

Circular waveguide

A circular waveguide is perfectly symmetrical and, as with a rigid guide, the inside must be free of imperfections such as dents and foreign material. A circular guide can handle circular, horizontal or vertical polarizations. The only difference is the way energy is fed and removed from the guide. It is preferred for long runs because it has less attenuation than either a rigid or an elliptical waveguide. It also has a cut-off frequency limit the same as rectangular or elliptical. One of the great advantages is where dual polarization is required for a given path, such as one polarization for analog and the other digital, two regular waveguide runs are eliminated. Or, in the case of AT&T, one frequency is horizontal and the adjacent channel is vertical. This cross-field approach yields about 30dB of isolation between polarizations. Where both ends of the waveguide have circular-to-rectangular transitions, it is considered a "closed system." An open system is one in which there is only rectangular transition at the feed point, the antenna having no transition. The horn or the cornucopia-style of antenna that AT&T used is an example of an open system.

A circular waveguide operated

below the cutoff frequency is used in the older signal generators for attenuation. The Measurements Corp. signal generator used the circular waveguide below the cutoff frequency by incorporating a small loop as the injector, and another loop

for pickup that was transported along the waveguide in a linear fashion. The linear distance was calibrated in decibels of attenuation. Attenuation of several orders of magnitude can be obtained using this method. ■

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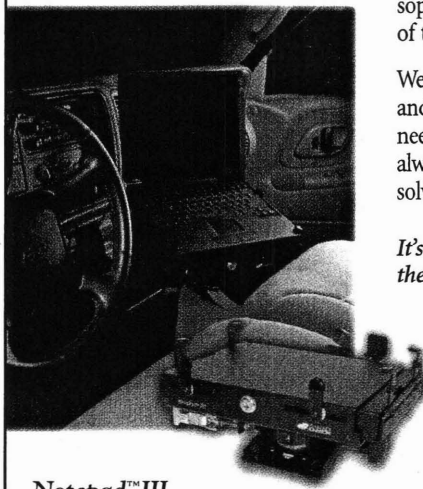
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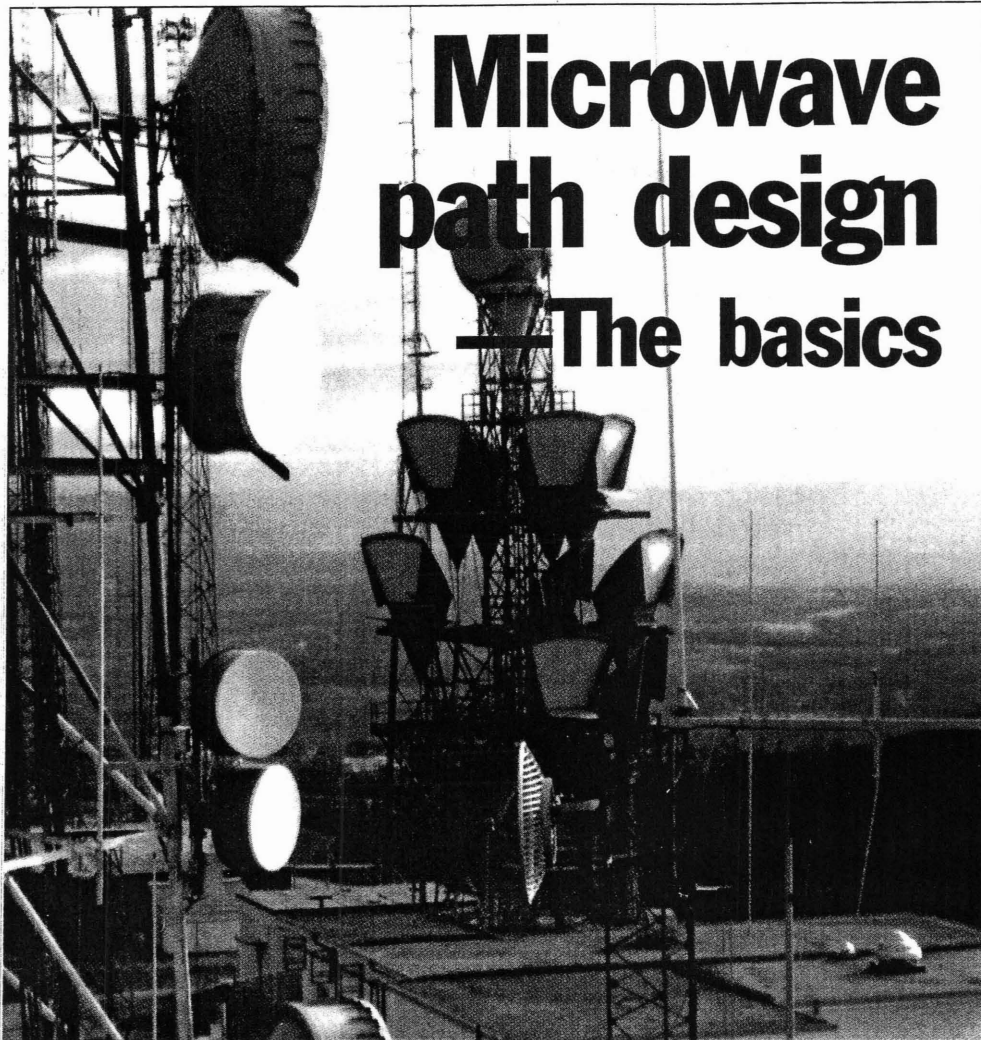
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Microwave path design

—The basics



Part 4—This series of four articles is a primer on some of the propagation-related basics of microwave path design. This last installment discusses path reliability.

By Jeff Ashley

Preceding installments in this series have described basic considerations:

- Determining whether a proposed path is "line-of-sight."
- Evaluating path clearances with regard to refractive effects.
- Evaluating path clearances with regard to Fresnel zones.
- Considering path reflections.
- Deriving a power budget and fade margin.

This concluding article covers path reliability.

Path reliability

Usually, microwave links are designed to meet a specific reliability factor. Reliability may also be termed "availability"

and is expressed as a percentage. It represents the percentage of time the link is expected to operate without an outage caused by propagation conditions. The Bell standard for short-haul propagation reliability is 99.995% (minimum), while requirements for high-capacity, long-haul may be 99.9999%. "Unavailability" or the probability of an outage (because of propagation conditions) is often referred to and, if expressed in percentage, the value is determined by subtracting the availability (expressed as a percentage) from 100. For example, a short-haul reliability (availability) factor of 99.995% can also be expressed as an unavailability factor of 0.005%.

Availability and unavailability are referenced to a year. In other words, if the unavailability is 0.001%, it could also be

stated there will be:

$$365.25 \text{ (days / yr)} \times 24 \text{ (hr / day)} \times 60 \text{ (min / hr)} \times \frac{0.001}{100} \\ = 5.26 \text{ minutes of unavailability (outage) per year.}$$

The propagation-related unavailability is derived from Equation 1 below. This equation is used as an approximation for the one-direction effects of multipath fading with various terrain and atmospheric conditions.

$$P_m = a \times b \times 2.5 \times 10^{-6} \times f \times D^3 \times 10^{-F+10} \quad [\text{Eq. 1}]$$

where

P_m = Outage probability (unavailability) due to multipath fading.

D = path length (in miles).

f = frequency (in GHz).

F = fade margin of the path (in dB).

a = one of the following generalized terrain factors:

4: very smooth (including water).

1: average, some roughness.

$1/4$: very rough.

b = one of the following generalized climate factors:

$1/2$: hot, humid.

$1/4$: normal, temperate.

$1/8$: very dry.

(Note that the result of this equation must be multiplied by 100 in order to be expressed as a percentage.)

Another factor that can affect path reliability is rain attenuation, particularly past 10GHz. At these higher frequencies, the wavelength becomes short enough that rain droplets (depending on their size and shape) can present themselves as obstructions to the path. This occurs when the diameter of the droplets approach $1/4$ wavelength of the frequency being used. The maximum instantaneous intensity of the rainfall and the percentage of the total path within the rain cell are other factors to be considered.

Because rain rates differ by geographic location, statistics for the particular area of interest must be obtained. Commercial path design software typically has this information in tables, requiring the user simply to select the geographic location of the proposed path. The values for that area are then accessed. The National Weather Bureau also maintains this information for various cities throughout the United States.

Because rain rates differ by geographic location, statistics for the particular area of interest must be obtained. Commercial path design software typically has this information in tables, requiring the user simply to select the geographic location of the proposed path. The values for that area are then accessed. The National Weather Bureau also maintains this information for various cities throughout the United States.

Increasing path reliability

One method of increasing path reliability is to increase the fade margin. If the

Ashley is a communications technician in Los Angeles.

design engineer dictates a particular fade margin requirement from the start, the power budget can be determined in a particular sequence. Knowing the noise threshold of the receiver being used (-dBm) and the required fade margin, the nominal signal level into the receiver can be determined (noise threshold [-dBm] plus fade margin [+ dB] = nominal receive level [-dBm]). The difference in decibels between the transmitter RF output level (+ dBm) on one end of the path and the desired receive signal level (RSL) on the other end of the path will be the maximum allowable net path loss. Because path distance and antenna heights are known, free space and atmospheric losses, along with all hardware losses on both ends, can be determined and will exceed this allowable net path loss by some amount (measured in decibels). The amount by which it is exceeded will be the necessary combined gain of the antennas on each end of the path. To achieve this required gain, proper antenna reflector diameters must be selected.

Parabolic ("dish") antennas typically have gain efficiencies of 55% to 65%. Using an efficiency of 55%, the gain of a parabolic antenna can be calculated as follows:

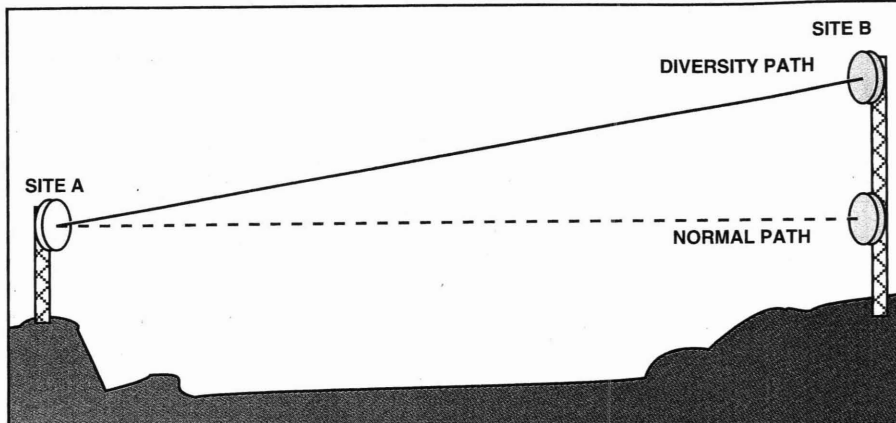


Figure 1. With adequate spacing between antennas, there should be little correlation between paths with regard to fading. *36dB e 280m*

$$G = 20 \log_{10} B + 20 \log_{10} F + 7.5 \quad [\text{Eq. 2}]$$

where

G = gain over an isotropic antenna.

B = parabola diameter in feet.

F = frequency in GHz.

After a proposed microwave path has been evaluated with regard to reliability, the calculations may show that it will not meet the required standards. In such situations, additional techniques known as *diversity* methods may be employed.

One such technique is space diversity, which involves the addition of another re-

ceive antenna, separated in distance from the first, as shown in Figure 1 above. An additional receiver and associated switching/combining equipment will also be needed. The additional "diversity" antenna is usually mounted on the same tower structure, yet vertically separated from the site's original receive antenna. This diversity antenna is connected to the diversity receiver. This additional equipment essentially creates another microwave path between the two sites. Receiver switching and combining equipment then senses which receiver has the troubled signal (if fades occur) and switches that receiver "off-line."

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One drawback of space diversity is increased cost. Tower heights must be greater to achieve sufficient vertical antenna separation. In some cases, the existing tower structure might have to be retrofitted to handle the additional wind loading created by the diversity antenna. The diversity antenna, receiver and combining and switching equipment create additional cost.

When spacing is adequate between antennas, there should be little correlation between the two paths with regard to propagation characteristics. Improvement

in reliability comes from the reduced probability that both paths will be adversely affected by fading at the same time.

In general, more vertical spacing between antennas offers less path correlation and better path reliability. The rule of thumb for antenna spacing is as follows: 60-80 feet at 2GHz, 45-50 feet at 4GHz, 30-40 feet at 6GHz and 25-30 feet at 11GHz.

The following modified form of Arvids Vigants' fade reduction factor equation can be used to estimate the improvement offered by space diversity:

$$I_{sd} = \frac{7.0 \times 10^{-5} \times f \times s^2 \times 10^{F+10}}{D}$$

[Eq. 3]

where

I_{sd} = improvement factor (space diversity).

f = frequency in GHz.

s = vertical antenna spacing, in feet, between centers.

D = path length in miles.

F = path fade margin (if one antenna's path has a smaller fade margin, use that figure).

Frequency diversity is another method used to increase path reliability. Microwave transmitters operating on two frequencies (with a typical in-band diversity spacing of about 2%), and sometimes in two frequency bands (called crossband diversity), are used to transmit the same information to separate receivers at the other end of the path.

Reliability improvement comes from the reduced chances of fading occurring on both frequencies (or frequency bands) at the same time.

Frequency diversity does not require the vertical antenna spacing used in space diversity, and therefore reduces tower cost. It does, however, require the use of more spectrum because it uses two sets of frequencies.

Because available frequencies are in limited supply, the FCC disallows the use of frequency diversity except under certain conditions. Test comparisons between space and frequency diversity indicate that improvements realized with space diversity are generally greater than those achieved with frequency diversity.

A general formula to calculate the improvement factor given by frequency diversity is as follows:

$$I_{fd} = K \times \left(\frac{\Delta f}{f} \right) \times 10^{F+10}$$

[Eq. 4]

where

I_{fd} = frequency diversity improvement factor.

$\frac{\Delta f}{f}$ = frequency diversity spacing (percent $\div$ 100).

F = fade margin.

$K = 1/2$ for 4 GHz band.

$1/4$ for 6 GHz band.

$1/8$ for 8 GHz band.

$1/12$ for 12 GHz band.

(The values of K shown above for the 8GHz and 12GHz bands are GTE-Lenkurt engineering estimates based on work from

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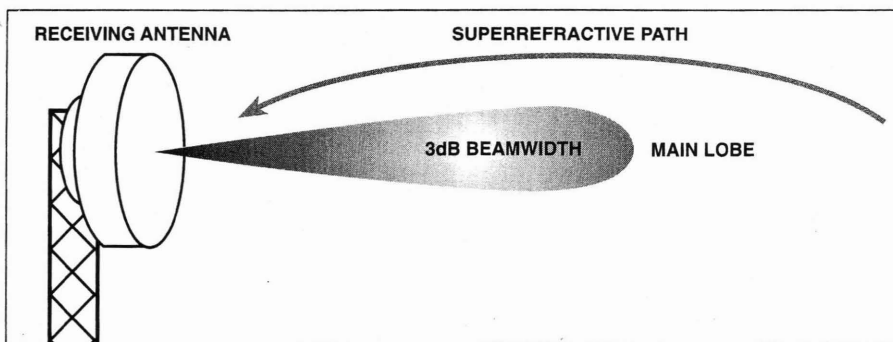
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The increased vertical angle of arrival under superrefractive conditions may cause the incoming wavefront to fall outside the beamwidth of the antenna. This is called 'antenna decoupling' and can cause a significant reduction in receive levels.

W.T. Barnett that dealt strictly with the 4GHz and 6GHz bands.)

Neither space diversity nor in-band frequency diversity provide improvement against rain attenuation, although out-of-band frequency diversity (such as 6GHz/12GHz) does.

Under superrefractive atmospheric conditions, negative values of K can exist where the path bows skyward at mid-path and comes downward at the ends. This will create a wavefront that will arrive at the antenna with a greater vertical angle than normal (see Figure 2 above). Because microwave antenna beamwidths can be small, this increased vertical arrival angle may be outside the antenna's half-power

beamwidth. The received signal level realized may be sufficiently low to cause noise problems or even an outage. This has been termed "antenna decoupling." A method of reducing this effect is to use a special *angle diversity* feed horn arrangement that essentially creates two vertically separated main lobes, so that if the signal level from the lower (normal) main lobe diminishes sufficiently because of the wavefront's increased arrival angle, the other (upper) lobe will be better oriented. Because most fades are due to reflections and multipath effects rather than this anomalous superrefractive condition, angle diversity offers only a limited improvement factor over that of a non-diversity path under typical path conditions.

Hybrid diversity is where both frequency and space diversity are employed simultaneously. However, most of the improvement comes from the effect of the space diversity.

The following equation is used when calculating fade probabilities with various methods of diversity:

$$U_{div} = \frac{P_m}{I}$$

[Eq. 5]

where

U_{div} = unavailability or probability of outage (using some diversity method).

P_m = annual unavailability (non-diversity) due to multipath (Eq. 1).

I = diversity improvement factor (Eq. 3 and Eq. 4).

In multiple-hop microwave systems, the overall reliability figure (expressed as a percentage) is obtained by taking the sum of the unavailability figures (as a percentage) for each hop, and subtracting that figure from 100.

Conclusion

Many other considerations go into the complete design of a microwave path and communications link. Additionally, each of the points discussed can be taken in much greater depth. Volumes of information are available to the interested reader from many of the large communications companies, both past and present.

This series of articles has offered a glimpse into some of the more fundamental propagation-related design considerations and will be of value to those just getting their feet wet in the microwave communications field. ■

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Acknowledgments

A special thanks to Cesar Llanes for answering so many questions through the years.

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Microwave path design: The basics

Part 2 —This four-part primer on propagation-related basics, for those learning about microwave communications, continues with a discussion of path clearances and Fresnel zones.

By Jeff Ashley

In Part 1 of this series (*MRT* May), two of the basic propagation-related considerations in the microwave path design process were discussed:

- determining whether a proposed path is "line-of-sight."

- evaluating path clearances with regard to refractive effects.

Remaining fundamental design components to be discussed are:

- evaluating path clearances with regard to Fresnel zones.

- considering path reflections.

- deriving a power budget and the fade margin.

- path reliability.

This article will briefly cover item three, path clearances and Fresnel zones. The remaining items will be dealt with in the con-

cluding parts of this series.

Evaluating path clearances—Fresnel zones

If one were to look directly into the center beam of a microwave path (gazing directly into the dish), the Fresnel zones could be defined as an infinite number of circles centered around the center beam, each having a larger radius than the previous one, as shown in Figure 1 below. The Fresnel zones get closer as they expand further away from the center. The radii of these zones reach a maximum at mid-path and diminish on either side of mid-path.

From a side view of the path, as shown in Figure 2 on page 30, one can see the creation of ellipsoids between path endpoints, centered around the main beam. Each ellipse represents the boundary of a particular Fresnel zone.

Fresnel zones are numbered from the

center of the path out. The boundary of the first radius out from the center represents the first Fresnel zone, which contains all wave components whose distance traveled to the far-end is exactly $1/2$ wavelength (λ) greater than the direct ray. The boundary of the second radius out from the center represents the second Fresnel zone, which contains all wave components whose distance traveled to the far-end is exactly 1λ greater than the direct ray. Each integer increase of Fresnel zone number represents an increase of $1/2\lambda$ to the distance traveled to reach the other end of the path.

The equation that determines the radius of the first Fresnel zone (in feet) at any particular distance along the path is:

$$F_1 = 72.1 \sqrt{\frac{d_1 \times d_2}{f \times D}}$$

where

F_1 = the first Fresnel zone radius (ft.).

d_1 = distance (mi.) from one endpoint to the point being considered.

D = total path length (mi.).

$d_2 = D - d_1$.

f = frequency (GHz).

The radius of any Fresnel zone number can be calculated by the following equation once the radius of the first Fresnel zone has been determined:

$$F_n = F_1 \sqrt{n}$$

where n = the Fresnel zone number being evaluated.

When examining path profiles, if one wishes to know what Fresnel zone is represented by a particular clearance (in feet), the

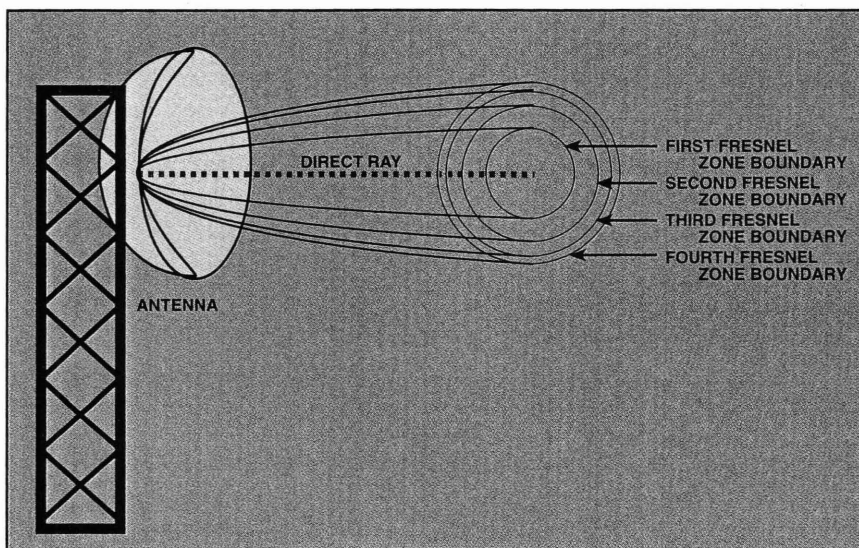


Figure 1. Fresnel zones could be viewed as an infinite number of circles centered around the direct ray, each circle having a larger radius than the previous one. Fresnel zones get closer together as they expand out further from the center.

Ashley is a communications technician in Los Angeles.

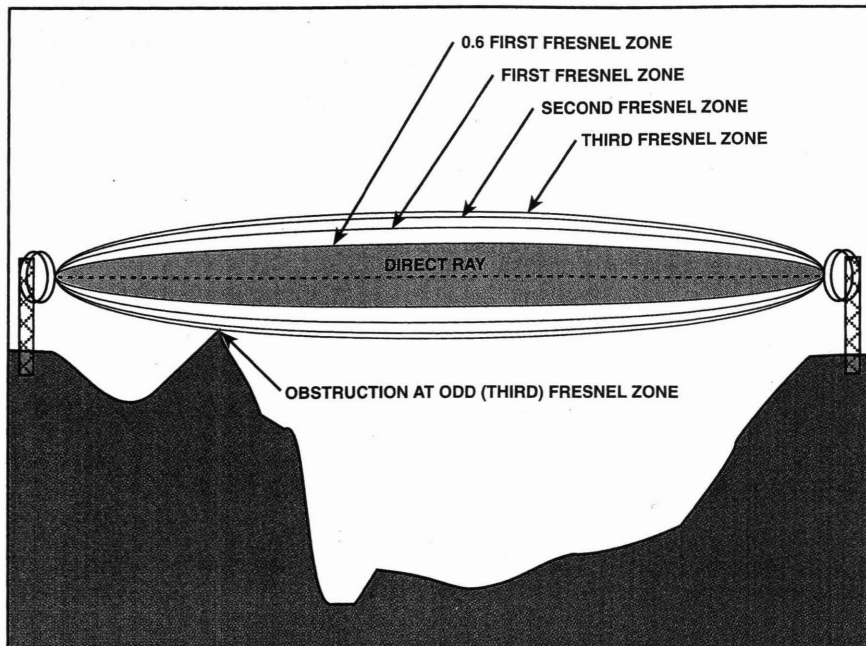


Figure 2. The Fresnel zone radii decrease with distance from mid-path. The Fresnel zone boundaries are defined by ellipsoids around the direct ray. Here, a path obstruction is shown with third Fresnel zone clearance. Because this is an odd Fresnel zone, signal energy reflected by the obstruction surface will be in phase and will add with the energy from the direct ray as it arrives at the far-end receiver. How much energy is reflected depends on the coefficient of reflectivity of the obstructing surface.

following form comes in handy:

$$\text{Fresnel zone \#} = \frac{\text{clearance}^2 (\text{ft.}) \times f \times D}{72.1^2 \times d_1 \times d_2}$$

In general, if the path is line-of-sight with adequate clearance, the loss in RF signal between one end's transmitter and the other end's receiver will merely be that of free space, as described by the equation:

$$FSL (\text{dB}) = 96.6 + 20 \log_{10} F + 20 \log_{10} D$$

where

$FSL (\text{dB})$ = Free Space Loss (dB).

F = Frequency (GHz).

D = Distance (mi.).

As indicated in Figure 3 on page 32, if a path has less than 0.6 first Fresnel zone clearance, there will be more loss than merely that of free space, even though a line-of-sight condition may exist. With odd Fresnel zone clearances something less than free space losses may be realized, while with even Fresnel zone clearances losses may be quite significant. This is because of the phase relationship between the direct and reflected



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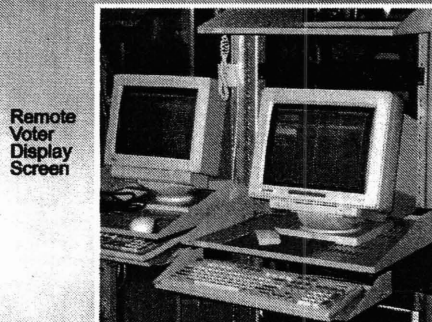
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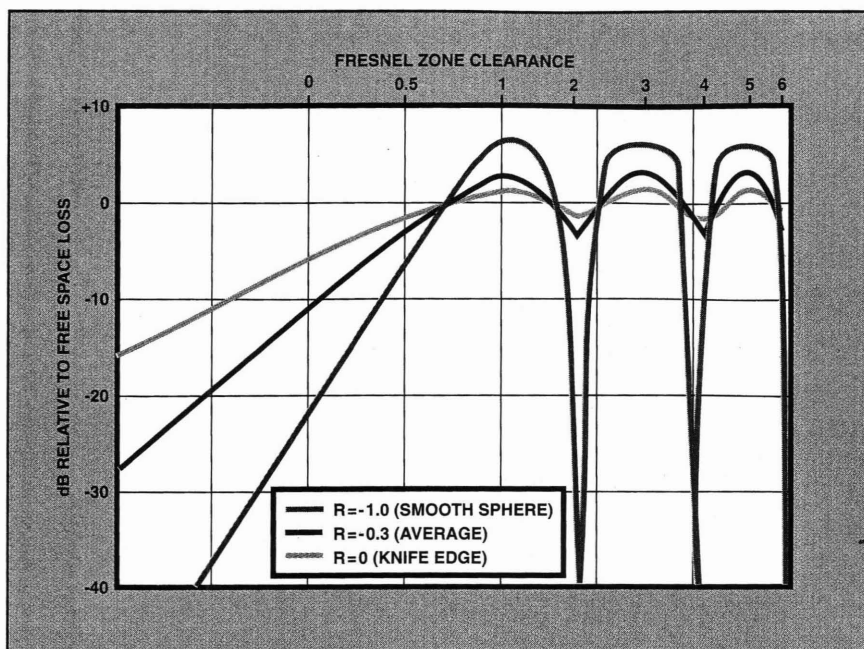


Figure 3. Deep fades result when out-of-phase reflected energy combines with the incident signal at the receiver, increasing the cancellation of the RF receive signal.

signal at the receiving antenna.

When the signal is reflected, it undergoes a 180° phase lag (equivalent to $1/2\lambda$ delay)

referenced to the incident wavefront. (Although, as will be mentioned in Part 3 of this series, wave polarization and grazing angle

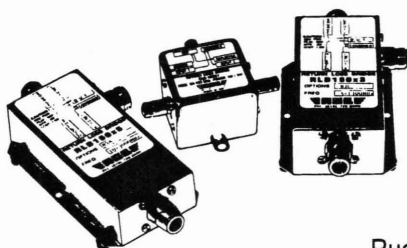
can affect this). If the reflective surface is located at the first Fresnel zone, the reflected signal will arrive at the receiver delayed by another $1/2\lambda$, for a total delay of 1λ . Signals reaching the receive antenna are therefore in phase and additive.

If the reflective surface is located at the boundary of the second Fresnel zone (which by definition, delays the signal by two $1/2$ wavelengths) the wavefront, when reflected, reaches the receiving antenna delayed by a total of 1.5λ . This means it will be out of phase relative to the incident wavefront, and signals reaching the receiver will cancel to some extent.

This effect repeats for each odd and even Fresnel zone, which explains the slight rises and very deep nulls in signal level for the even and odd Fresnel zone clearances (as shown in Figure 3). Raising one or both antennas is one method of altering the Fresnel zone clearance. As the clearance increases to higher order Fresnel zone numbers, the intensity of this problem diminishes because the magnitude of the coefficient of reflection with regard to the reflecting surface diminishes.

Paths are usually designed to meet specific Fresnel zone clearance criteria for one or more values of K . "Heavy route"

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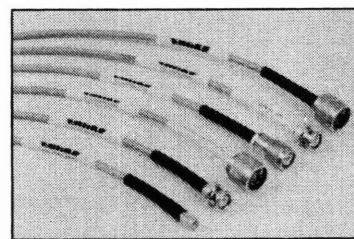
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- ☆ Listen before you transmit on a shared channel. Learn how to disable the squelch to listen for other users.
- ☆ Use your call sign whenever you use the radio.
- ☆ If you use a scrambler for privacy on a shared channel, sign off "in the clear" so other users know you are finished.
- ☆ Have your radio serviced periodically to be sure it is on frequency.
- ☆ Keep communications brief to promote harmony with other users who share the channel.
- ☆ Use the minimum transmitter power and antenna height you need to cover your area. That might be less than the power and antenna height specified on your license.

System operators:

- ☆ Post the name and 24-hour phone number of the site manager at the access point for each antenna site.

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requirements dictate that there must be at least 0.3 first Fresnel zone clearance when $K = 2/3$, and 1.0 first Fresnel zone clearance when $K = 4/3$, whichever is greater. "Light route" requirements dictate at least 0.6 first Fresnel zone clearance plus ten feet when $K = 1$.

Another important consideration here is what type of reflecting surface is being evaluated. In general, three categories of surfaces are defined, each with its own reflection coefficient: $R = 0$ (a sharp or "knife" edge), $R = -0.3$ ("average" roughness) and $R = -1.0$ (a smooth spherical surface). (The reflection coefficient will be described in more detail in the third article of this series.) The type of surface makes a big difference with regard to attenuation on the even Fresnel zone clearances because obstructions with larger, smoother surface areas reflect more energy. With more out-of-phase reflected energy combining with the incident signal at the receiver, greater cancellation of RF receive signal will be realized, resulting in deep fades as shown in Figure 3.

The next article in this series will briefly discuss path reflections, the power budget and the fade margin.

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Microwave path design series

Part 1: "Microwave path design: The Basics," May 1998.

Part 2: "Microwave path design: The Basics," June 1998.

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Microwave path design: The basics

Part 3—This series of four articles is a primer on some of the propagation-related basics of microwave path design. This month's installment explains the effect of path reflections on fading, how to derive a power budget and the fade margin.

By Jeff Ashley

In the first two parts of this series (MRT May and June), the microwave path design process was described in terms of:

- determining whether a proposed path is "line-of-sight."
- evaluating path clearances with regard to refractive effects.
- evaluating path clearances with regard to Fresnel zones.

This section covers path reflections and how to derive a power budget and fade margin. The concluding installment will examine path reliability.

Considering path reflections

A point of reflection will exist somewhere along the length of the path where (with regard to the reflecting plane and antennas at both ends) the angle of incidence equals the angle of reflection (see Figure 1 below). If antennas at each end of the path are at the same height above a flat, reflecting surface, the reflection

point will be located halfway between the two. However, if antenna elevations are different, the reflection point will be closer to the lower antenna and will move along the length of the path, toward one endpoint or the other as K varies ("K" was defined in Part 1 of this series). If sufficient out-of-phase signal from the reflection point is received at the far-end receiver, catastrophic fading may occur.

In some cases the reflection point will be blocked naturally by terrain features. Manipulation of antenna heights is a method of moving the reflection point to a different spot along the length of the path so that it can be blocked by an obstruction or perhaps so that it may fall on a less reflective surface. Rough surfaces do not reflect as well as smooth, flat surfaces, such as water.

The magnitude of the reflection coefficient (amount of "reflectivity") also increases as the angle of incidence decreases. Because of this, short paths with endpoints at a considerable height above the reflecting surface will be affected less

than those with endpoints barely higher than the reflective surface itself (a microwave link between two offshore oil platforms, for example). The magnitude of the reflection coefficient can vary from 0 to -1.0, the latter corresponding to 100% reflection. The negative sign refers to a 180° phase reversal at the point of reflection.

There is an advantage to using vertical polarization when the grazing angle on a reflective surface is a couple of degrees or greater. The magnitude of the reflection coefficient with vertical polarization is much less than that of horizontal polarization. Less reflected signal will be present at the receive antenna on the other end of the path. In fact, with a grazing angle of 6.5°, the magnitude of the reflection coefficient for a vertically polarized wave is only about -0.08 compared to about -1.0 for horizontal polarization. The magnitude rises as the grazing angle increases above 6.5° until it finally meets the same value of the horizontally polarized signal (at 90°).

The phase change (or lag) at the point of reflection for a vertically polarized signal is also significantly affected at low grazing angles as compared to the practically uniform 180° phase reversal for a horizontally polarized signal. As the angle increases from 1° to 30°, the phase change of a vertically polarized signal drops quickly from about 180° to 4°. With only a 4° phase lag, the reflected signal that reaches the receive antenna on the other end of the path would have more of an additive, rather than canceling effect.

The distance to the reflection point can be calculated using the equation

$$d_1 = nD \quad [\text{Eq. 1}]$$

Ashley is a communications technician in Los Angeles.

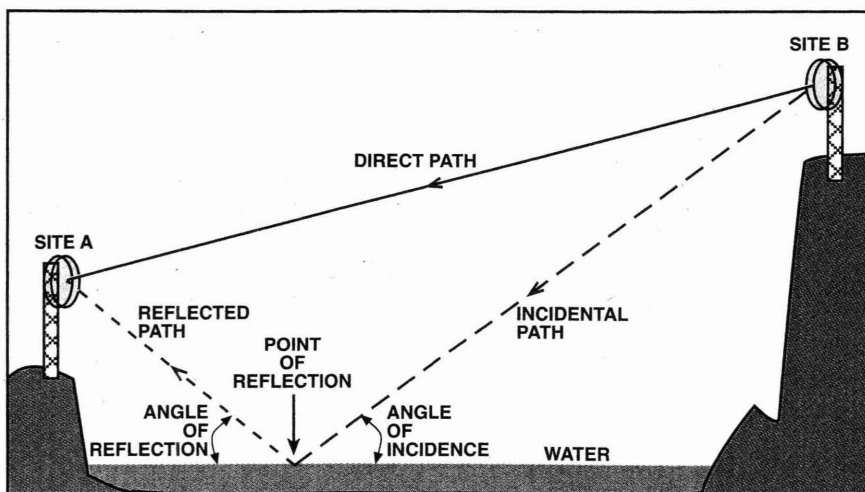


Figure 1. At the point of reflection, the angle of incidence equals the angle of reflection.

where (for $K = \infty$ only):

$$n = \frac{h_1}{h_1 + h_2}$$

and

h_1 = the elevation of the lower antenna above the reflecting surface.

h_2 = the elevation of the higher antenna above the reflecting surface.

d_1 = the distance (mi.) to the reflection point from the h_1 end.

D = the total path length.

Because calculations for n , where $K = 2/3$ and $K = 4/3$, get a bit involved, estimated values can be obtained from graphs offered in R.F. White (see references).

As refractive conditions change, the degree of bowing that occurs on the wavefront will change. As the amount of bowing changes, the location of the reflection point will vary in relative distance

between the two path endpoints. In designing a path, it is useful to analyze the overall range of movement (or distance of travel) that can occur for this reflection point. The design engineer must determine whether (in its range of movement) it may wander onto a highly reflective surface (such as a body of water) that lies between path endpoints. In such an undesirable situation, measures might be taken to move the reflection point elsewhere by changing antenna heights, or by attempting to block the reflection.

To determine the practical limits of travel for the reflection point, two values of K are typically used that would describe limits of refractive conditions. One limit is when refractive conditions correspond to a value of $K = \infty$. The other extreme limit will be the value of K that would bend the path downward to the point of grazing the ground. Beyond this grazing point, no reflection can exist because the path itself is being obstructed by the earth.

The value of K that corresponds to "grazing" can be calculated from the equation:

$$K = \frac{1}{1.5(X + Y + 2\sqrt{XY})} \quad [\text{Eq. 2}]$$

where

$$X = \frac{h_1}{D^2} \quad \text{and} \quad Y = \frac{h_2}{D^2}$$

(h_1 , h_2 and D are already defined in Equation 1).

The distance to the grazing point is calculated from Equation 1, where:

$$n = \frac{1}{1 + \sqrt{\frac{Y}{X}}}$$

Deriving a power budget

A path power budget is nothing but an itemized list of all system losses and gains (in decibels) from the transmitter on one end of the path to the receiver on the other end, and everything in between.

For example, if starting with the transmitter on one end of the path, one would note transmitter output power (+dBm), then subtract all the losses in subsequent connectors and circulators external to the radio assembly, along with waveguide losses (-dB) until the antenna is reached. The gain of the antenna would be added to that value (minus the loss in the radome, if used) to give the resultant effective radiated power (ERP). This is the available RF signal that will be propagated through the atmosphere to the other

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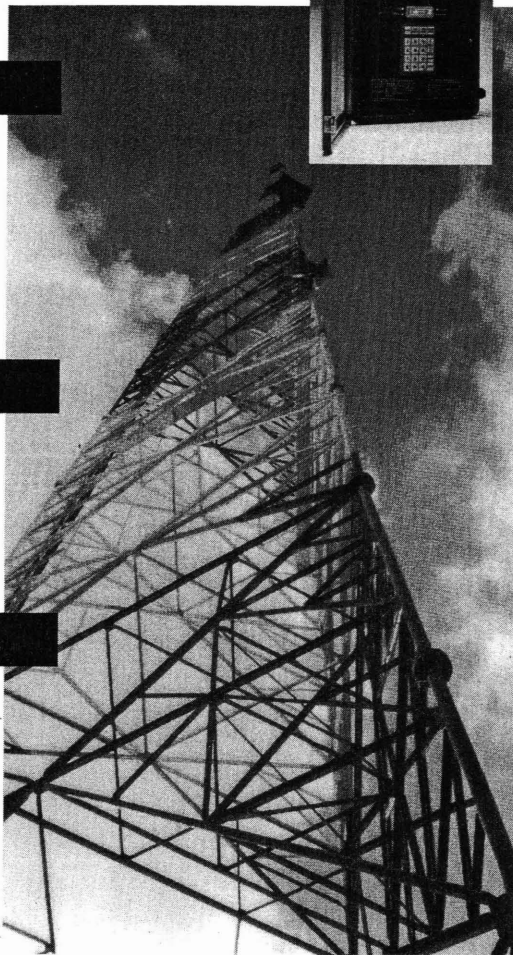
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end of the link. As the signal makes its way through the atmosphere to the receiving site, free space loss (-dB) (the formula for this was given in Part 2, *MRT* June), in addition to atmospheric losses (-dB), would then be subtracted from the ERP value.

The term atmospheric loss or "absorption" pertains to effects that atmospheric gases, primarily oxygen and water vapor, have on microwave propagation. These

effects are calculated separately from free space losses. Attenuation figures are directly proportional to path length and increase with frequency (0.0002 dB/mi. at 2GHz, 0.002 dB/mi. at 8GHz and almost 0.2dB/mi. at 20GHz).

At the receiving end of the link, the considerations are the same as on the transmitter end, only in reverse order (radome loss; antenna gain, waveguide, connector and hybrid/coupler losses), with

the remaining signal level being applied to the receiver (-dBm).

Fade margin

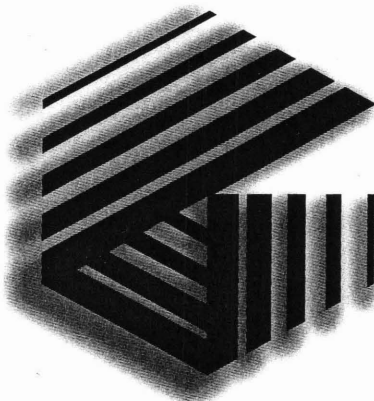
The receiver manufacturer will offer a noise threshold specification (expressed in -dBm) which, simply stated, is the sensitivity of the receiver. It is considered to be the lowest receive signal level at which the receiver will still be considered "unusable." There are variations in the definition of the term "unusable," which means the use of different standards in determining how a particular receiver threshold specification will be obtained. For example, on most industrial analog microwave systems the practical threshold of the receiver is considered the absolute RF signal level (-dBm) injected into the receiver that will establish a 30dB (flat) signal-to-noise ratio (58dBmC0) as measured in a discriminated baseband noise slot. The telephone companies, however, have used a 33dB (flat) signal-to-noise ratio (55dBmC0). (The baseband frequency of this noise slot will typically be specified by the equipment manufacturer and is dependent on the particular channel loading of that equipment. It is usually assigned somewhere above the highest occupied frequency slot on the baseband. Because noise effects are worst on the high end of the baseband, this slot would offer a worst case scenario for the signal-to-noise measurement.)

The threshold specification on digital microwave receivers makes reference to the bit-error rate (BER) of the recovered data rather than to a discriminated analog noise measurement. The old Bell standard for digital systems is the RF signal level injected into the receiver that will produce a BER that exceeds 10^{-3} . This is typically referred to as the "outage" threshold. Often, however, a BER of 10^{-6} (referred to as the operating point threshold) is used because this is a more common T-1 level standard. (As a note, the difference between these two bit-error rates relates to a change in RF receive level of only 2dB-3dB. This is because BER performance on digital radios degrades rapidly after the 10^{-6} point.) As long as the BER is defined along with the threshold specification, no confusion over terms should exist.

On analog microwave systems, the difference between the receiver threshold value (-dBm) and the receive signal level (RSL) being applied to the receiver (-dBm) under normal path conditions is referred to as the fade margin (expressed in dB). This is a "safety margin" of excess signal that the path can fade before the receiver becomes "unusable" due to

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noise. The same general definition could be used for digital microwave systems, although in this case it might be termed "flat" or *thermal fade margin* (TFM). Other terms applicable to digital systems are *dispersive fade margin* (DFM), *adjacent channel interference fade margin* (AIFM), and *external interference fade margin* (EIFM). *Composite fade margin* (CFM) is a term used to describe the composite affects of TFM, DFM, AIFM and

EIFM. Typical receiver threshold levels for digital or analog commercial microwave receivers are usually in the ballpark area of -70dBm to -85dBm .

It should be noted that microwave receivers have a specified range of RF input signal level over which they should properly detect and decode modulated information. This means not only that there is a *minimum* signal level (receiver "threshold") but also a *maximum* signal level that can be applied

to the receiver input without expecting distortion or data errors in the recovered signal. This restricts the path designer from creating a *huge fade margin* by applying excessively high RF levels to the receiver under normal path conditions. Additionally, a path phenomenon referred to as "up-fading" can occur (particularly around the end of a period of severe multipath activity) where the microwave link receivers realize an extreme increase in RF signal level (relative to normal RSL). Up-fading is thought to be the result of an anomalous "lens" effect in the atmosphere. For such reasons, the design engineer should leave a margin of at least 10dB between normal path RSL and the receiver's upper RF input limit. The typical design RSL for a microwave receiver is about -35dBm .

The next and final article in this four-part series will discuss microwave path reliability.

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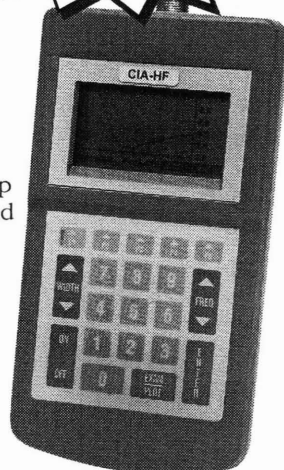
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Microwave path design: The basics

Part 1 — This four-part primer on propagation-related basics, for those learning about microwave communications, begins with "line-of-sight" determinations and the evaluation of path clearances with regard to refractive effects.

By Jeff Ashley

Microwave communications path design poses many challenges. In addition to *static gain and loss* considerations, *terrain* and *propagation dynamics* can play a large role in determining whether a proposed path will have the required signal levels, clearances and reliability.

This four-part series will only scratch the surface with regard to these considerations, and it is not intended to cover *all* aspects of path design or any one in great depth. The primary focus of the series is to serve as a primer on some of the propagation-related basics for those who may be new to the microwave communications field.

Given this disclaimer, the following tasks are some of the fundamental compo-

nents of microwave path design:

- ☐ determining whether a proposed path is "line-of-sight."
- ☐ evaluating path clearances with regard to refractive effects.
- ☐ evaluating path clearances with regard to Fresnel zones.
- ☐ considering path reflections.
- ☐ deriving a power budget and the fade margin.
- ☐ path reliability.

Part 1 of this series deals with the first two components; the others will be discussed in Parts 2, 3 and 4.

Determining line-of-sight

Although there are situations where microwave paths are designed to work without line-of-sight conditions, they would not be considered the "norm" for industrial micro-

wave systems and therefore will not be dealt with here. In most situations, if a prospective path is not line-of-sight, then an alternate route is considered.

Determining whether a path is line-of-sight can be partially accomplished with the aid of a *topographical map*. This type of map will show the various elevations along the length of the path between proposed endpoints. Plotting these elevations at intervals will produce a path profile showing terrain relative to the antenna elevations, as shown in Figure 1 below left. This graphical representation aids in determining not only whether a line-of-site condition exists between endpoints but also in measuring clearances between the main beam (center) of the path and the surrounding terrain.

Topographical information for many areas throughout the world is now available on CD-ROM. Various software applications have been developed to design microwave paths based on this information. Since this option can be a bit "pricey," many opt to use inexpensive topographical maps, which may be used with the same success.

When evaluating a proposed path, the *path profile* should be developed first. This will identify path obstructions from terrain features. A field survey should follow, which offers the necessary visual confirmation that the height of foliage or of man-made objects (which are not indicated on a topographical map) will not be located in or too near the proposed path.

If visual confirmation is difficult because of the distance between path endpoints, a mirror can be used to reflect the sun's light from one path endpoint back toward the other. If the mirror's flash of reflected sunlight is seen back at the opposite endpoint, the path is line-of-sight. This is referred to as "flashing" a path. Alternate methods

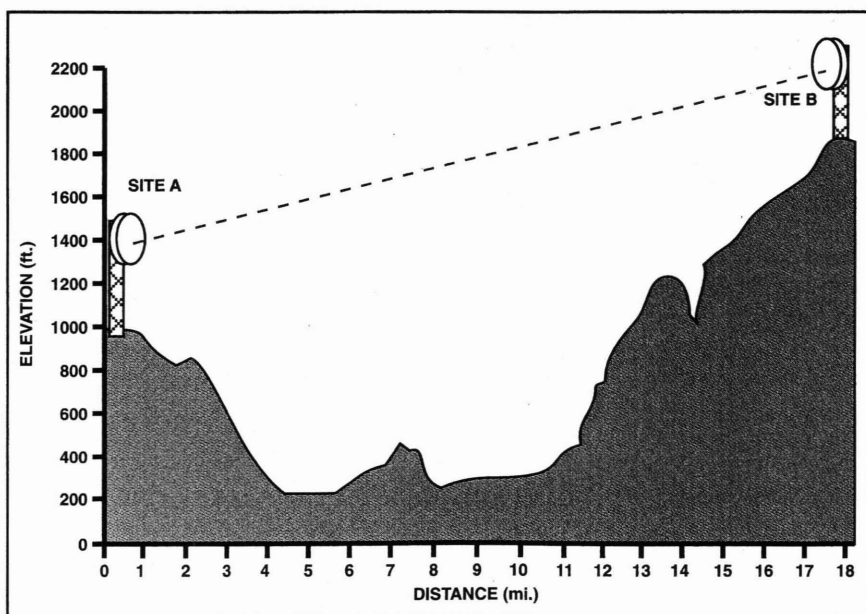


Figure 1. A typical path profile indicates terrain elevations relative to distance along the length of the path.

Ashley is a communications technician in Los Angeles.

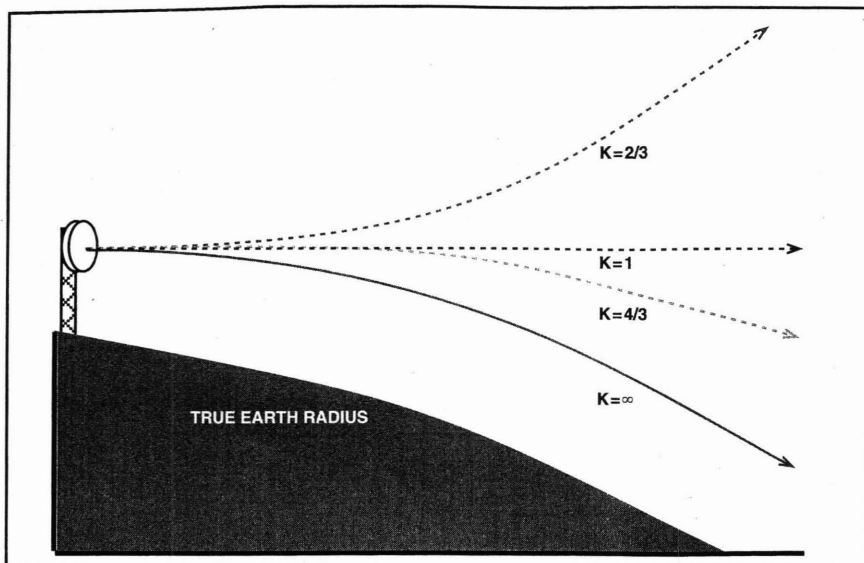


Figure 2. Where K equals infinity, the wavefront parallels the radius of the true earth. When no refractive effects are present ($K = 1$), no bending of the wavefront occurs.

might include the use of high-intensity search or strobe lights.

Evaluating path clearances—refraction

When the dielectric constant of the atmo-

sphere changes with height-above-ground, the refractive index will also change. This refractive variation in turn causes the propagating wavefront to effectively "bend." If a condition exists where the dielectric constant

of the atmosphere is constant with height-above-ground, no refraction will occur and the wavefront will travel in a straight line. This is not the "norm," however. Because of dielectric variations typically present, a wavefront usually will be refracted so that it will follow a path somewhere between a straight line and the true curvature of the earth (referred to as R). A wavefront propagating through the environment will be slowed by, and be effectively bent toward, an atmosphere that is more dense, such as one with a high water-vapor content (dense clouds or fog). That is why, rather than traveling in a straight line, the wavefront is normally bent earthward—which serves to extend its horizon.

Because atmospheric conditions are dynamic, the bending effect of the propagating wavefront will vary. To provide a simple model describing the path traveled by a wavefront for a particular refractive condition, engineers have developed a factor, K , where KR is considered the "equivalent" earth radius. This equivalent radius describes the bending of the wavefront relative to the true earth radius, as shown in Figure 2 on page 28.

The refractive index can change

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drastically with time. This means the micro-
wave beam between path endpoints will
"bow" to a greater or lesser extent. In fact,
it may "bow" in an upward or a downward
direction, depending on the value of K at the
time. It is therefore important to evaluate
clearances from path obstructions over a
wide range of K in order to determine
whether adequate path clearances are main-
tained under various degrees of refraction.
Often, three values of K are used in the cal-
culations. Two of the three values describe
the *limits* or boundaries of refraction that
might occur (although with anomalous con-
ditions, values exist outside these limits),
while the third describes what is considered
"normal" or expected. The value of $K = \infty$,
also known as "super-standard" atmosphere,
is one extreme where the wavefront follows
the true curvature of the earth. The other
extreme value of K typically used is $2/3$ and
is termed "substandard" atmosphere. It is
also a condition commonly referred to as
"earth bulge." The median value of $K = 4/3$
is used to evaluate the path under "normal"
atmospheric conditions in temperate
climates.

There are two methods of graphically
displaying the various *equivalent* earth radii

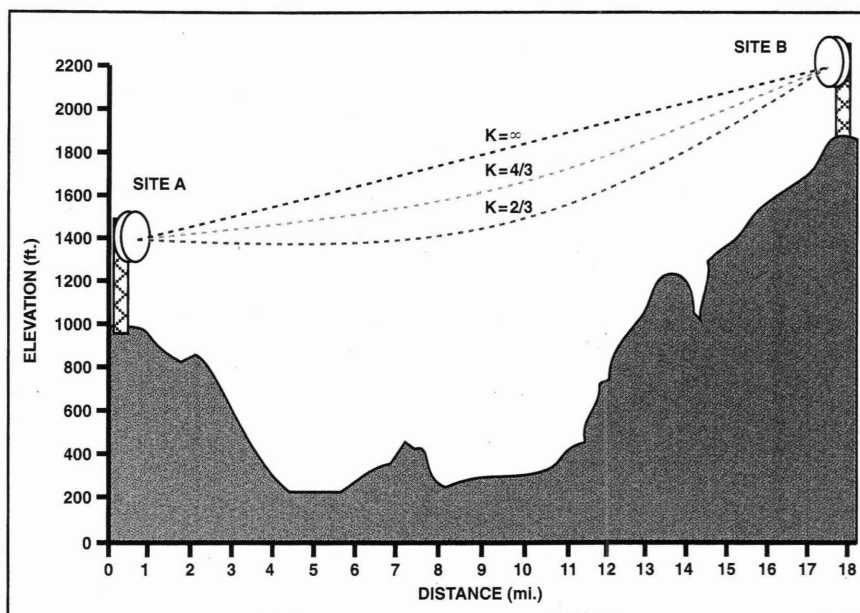


Figure 3. This is a 'flat-earth' profile. Terrain features become more of an obstruction consideration when $K = 2/3$.

with respect to the terrain between path
endpoints: the "flat earth" method and the
"curved earth" method. Figure 3 above is an

example of a "flat earth" path profile. It is
called "flat earth" because the horizontal axis
(or baseline) has been kept flat and

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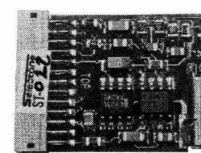
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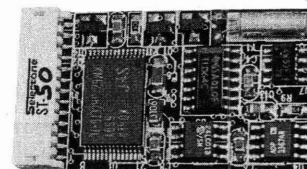
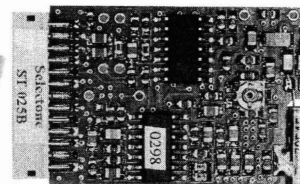
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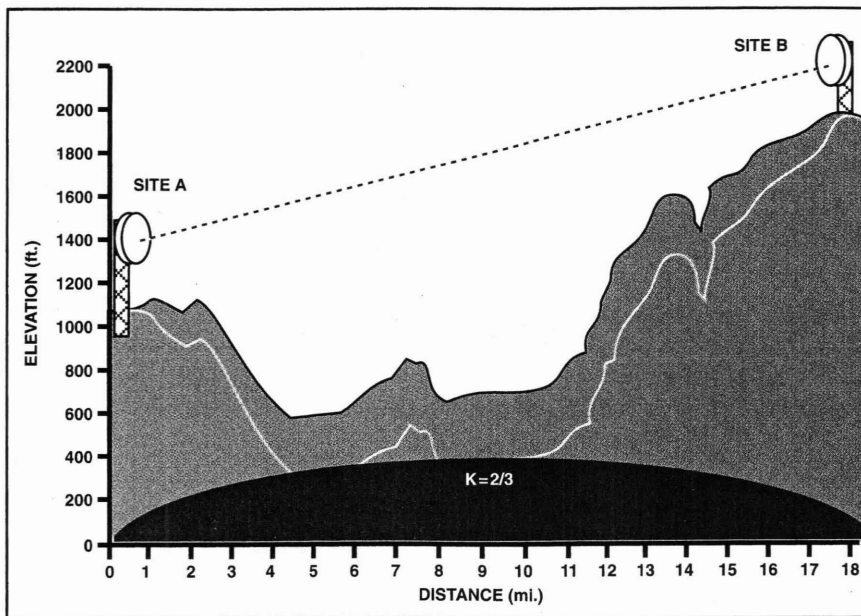


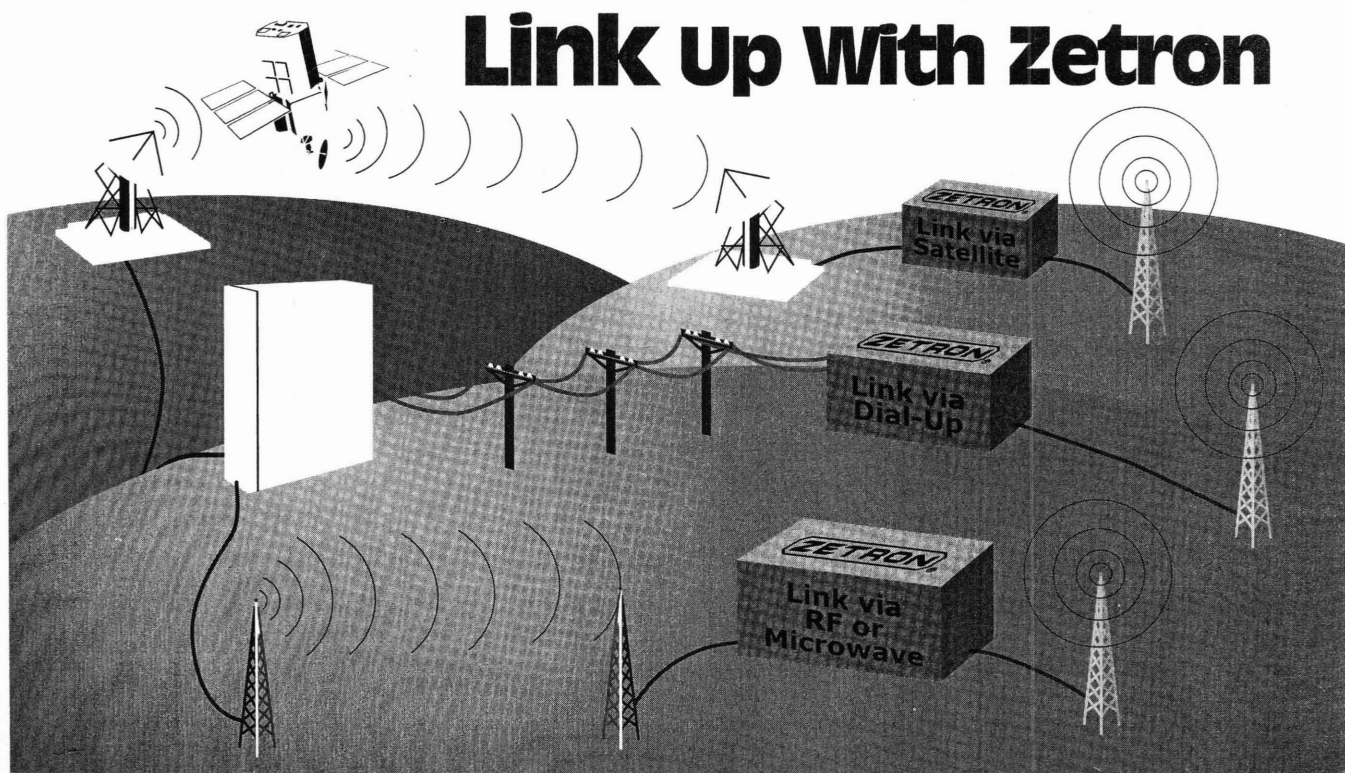
Figure 4. This is an exaggerated curved-earth profile where $K = 2/3$, with identical terrain features as shown in Figure 3. (Figure 3 elevations are represented by the yellow line.) Because the equivalent earth radius 'bulges up' in the center, the elevations of the terrain features increase, reducing path clearances. When $K = 4/3$, the earth bulge is less severe, resulting in greater path clearances. The greatest clearance is realized when $K = \infty$.

represents the earth's curvature as a flat line. The various equivalent earth radii represented by the different values of K are shown by bending the path beam line between antennas. The three lines shown in Figure 3 correspond to the refractive conditions where $K = \infty$, $K = 4/3$ and $K = 2/3$.

Elevations of terrain features between path endpoints are plotted using the flat baseline as a reference. The profile allows one to see if there is sufficient path clearance over the terrain for each value of K .

The curved earth approach, as shown in Figure 4 on the left, leaves the beam line between antennas straight and instead bows the horizontal baseline to correspond to one particular value of K . Elevation of all terrain features must be added to that of the bowed baseline reference. The straight beam line between antennas must now be checked for sufficient clearance from all terrain features for that value of K .

With the curved earth profile, the one limit of $K = \infty$ would be shown with the baseline flat. Here (as with the flat-earth approach) is the one "extreme" condition where the wavefront (shown as a straight line) follows



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the true curvature of the earth (also shown as a straight line). As the value of K gets smaller and reaches the other limit of $K=2/3$, the equivalent earth radius bulges upward more in the center, hence the term "earth bulge." As these profiles show, paths with values of $K = 4/3$ and $K = 2/3$ have less path clearance with respect to terrain features than when $K = \infty$.

The following equation can be used to calculate how much the horizontal axis (or

baseline) of a "curved earth" profile will bow upward at any point along the path for a particular value of K :

$$h = \frac{d_1 \times d_2}{1.5 \times K}$$

where h is the vertical variation in height (ft.).

d_1 is the distance (mi.) from one end of the path to the point being considered.

d_2 is the distance (mi.) from the point considered to the other end of the path.

With the "flat earth" profile, h is subtracted from the direct ray ($K = \infty$) reference line (for positive values of K) at various distances along the path in order to develop a graphical representation of the amount of wavefront bowing that will occur for the value of K under investigation.

With the "flat earth" profile, the wavefront, rather than the baseline, is shown "bowed," and the amount of bowing changes as the value of K changes. The distance between the bowed wavefront and the terrain features can be measured graphically to determine path clearances for each value of K .

Whichever profiling method is used ("curved earth" or "flat earth"), it can be seen graphically that smaller values of K result in greater bulging or "bowing." This in turn

... Smaller values of
 K result in greater
bulging or 'bowing.'
This in turn results in
less path clearance ...

results in less path clearance with regard to terrain features.

Because the microwave path has *radial dimension*, the clearances considered are not only those above and below the center of the path, but also to either side.

Initially, these graphical methods allow path clearances to be measured as distances, *in feet*. Once this information has been obtained, it is however, typically converted to a more meaningful design measurement, called the *Fresnel zone*.

The next article in this series will discuss path clearances with regard to Fresnel zones.

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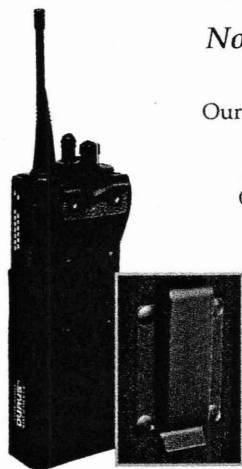
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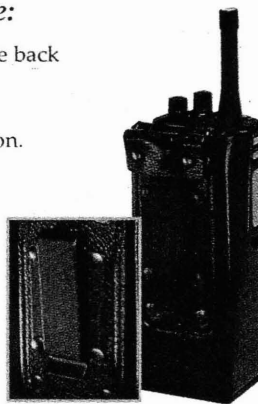


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Site noise vs. system noise figure

By Harold Kinley, C.E.T.

No matter how good the noise figure of individual components of a receiving system may be, it is the overall system noise figure that represents the bottom line in describing how well a receiving system will perform in the real world. To a large degree it is the ambient site noise in which the receiving antenna is placed that determines the overall system noise figure. In this column we will look at several receiving systems for a conventional narrowband FM ($\pm 5\text{kHz}$ deviation) system and compare the performance in high-, medium- and low-noise environments.

Antenna noise figure

Sometimes you will see the term *antenna noise temperature* used to describe the level of ambient site noise in which the antenna is placed. I prefer to use the term *antenna noise figure*. Figure 1 below shows a simple open-circuit resistor. Thermal noise voltage will appear across the resistor. The level of the noise voltage will depend on the resistance, temperature and bandwidth. The formula for noise voltage is (V_N):

$$V_N = \sqrt{4RkTB}$$

For 50W systems this can be reduced to

$$V_N = 14.14\sqrt{kTB}$$

where

k = Boltzmann's constant, or 1.38×10^{-23}

T = temperature in degrees Kelvin (K)

and

B = bandwidth in hertz.

Generally, 290°Kelvin, or about 62°F, is considered to be earth's temperature. The noise voltage appearing across the 50Ω resistor shown in Figure 1 is 0.1096μV at 290°K in a 15kHz bandwidth.

The equivalent noise input of a receiver can be determined by the following for-

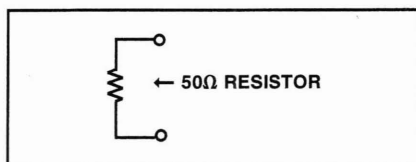


Figure 1. The open circuit thermal noise voltage across the 50Ω resistor is 0.1096mV

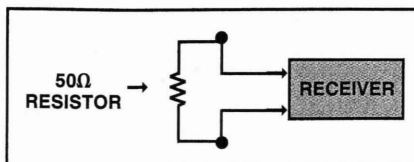


Figure 2. When the resistor is connected to the input of the receiver, the noise voltage from the resistor is reduced by one-half to 0.055μV, and the receiver with a noise figure of 0dB will have an equivalent noise input voltage of 0.055μV. This results in a total noise voltage of 0.0778μV at the receiver input.

mula: $N_R = 10 \log(N_B) + N_F - 174$ where N_R = equivalent receiver input noise in dBm, N_B = noise bandwidth in hertz and N_F = receiver noise figure in decibels. If the receiver noise figure is 0dB and the noise bandwidth is 15,000Hz, then the equivalent noise input to the receiver is -132.24dBm or 0.0546μV rounded to 0.055μV.

Now, refer back to the 50Ω open-circuit resistor. It had a noise voltage of 0.1096μV across the terminals. If we connect this resistor to a 50Ω receiver with a 0dB noise figure (with the equivalent of 0.055μV noise voltage input), what happens? (See Figure 2 above.) Because the resistor is terminated in 50Ω (receiver input impedance), the noise voltage across the resistor will divide in half to produce 0.055μV of noise voltage across the receiver input. Now the resistor is contributing 0.055μV of noise, and the input noise at the receiver is 0.055μV. Because the two noise voltages are non-coherent, the total or resultant noise voltage at the input to the receiver is found from the *root-sum-square* (RSS) of the two voltages.

Thus, the resultant noise voltage at the input to the receiver is:

$$\begin{aligned} N &= \sqrt{N_R^2 + N_{Rx}^2} \\ &= \sqrt{0.055^2 + 0.055^2} \\ &= \sqrt{0.00605} \\ &= 0.0778\mu V \end{aligned}$$

Thus, with the 50Ω resistor connected to the receiver input, the noise voltage has increased from 0.055μV to 0.0778μV, representing an increase of 3dB. The noise figure has increased by 3dB with the connection of the resistor to the receiver, or simply, the resistor has a noise figure of 3dB. If we substitute an antenna for the resistor the antenna will have the same effect on noise figure; that is, assuming that the antenna is placed in an environ-

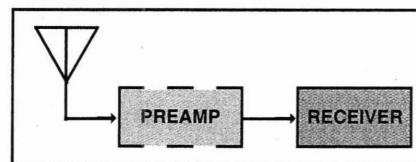


Figure 3. The basic receiving system is composed of the antenna, the preamplifier, transmission lines and the receiver.

ment where site noise is negligible. It would be hard to find such a place in the real world. The main point is to show that an antenna will exhibit a noise figure of 3dB at a minimum. Normally, the equivalent antenna noise figure will be *much higher* than 3dB.

System noise figure

Now, we will examine several examples of receiving systems to compare the effects of site noise with various amounts of line loss and with or without a tower-top preamplifier. Figure 3 above is our basic receiving system. The preamplifier (the dashed line) may or may not be in line for a particular series of data.

The receiver used in this table represents a sensitivity of approximately 0.25μV for a noise figure of 9dB. The first four lines of the table are for antennas with the best possible noise figure (3dB) representing a site with the lowest possible noise. Notice that the line loss affects the overall system noise figure almost on a decibel for decibel basis.

Lines 5-9 illustrate what happens when the antenna is at a site with moderate noise represented by a 10dB antenna noise figure. Notice that on line 5, for a 0dB line loss, the system noise figure is 12.3dB. On line 7, the line loss is increased to 4dB but only increases the system noise figure by slightly more than 2dB.

Lines 9-11 show some comparisons at a site where the site noise is high represented by a 20dB antenna noise figure. The overall system noise figure with a

(continued on page 50)

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Technically speaking

(continued from page 8)

Table 1. How site noise affects line loss and the use of a tower-top amplifier.

Line	N _{ANT} (dB)	PREAMP	L _{LOSS} (dB)	N _{RX} (dB)	N _{SYS} (dB)
1	3	NO	0	9	9.5
2	3	NO	1.5	9	10.9
3	3	NO	3.0	9	12.3
4	3	NO	6	9	15.1
5	10	NO	0	9	12.3
6	10	NO	2	9	13.3
7	10	NO	4	9	14.6
8	10	NO	6	9	16.1
9	20	NO	0	9	20.3
10	20	NO	3	9	20.6
11	20	NO	6	9	21.2
12	3	NO	0	9	9.5
13	3	YES	0	9	4.2
14	3	YES	3	9	4.6
15	3	YES	6	9	5.3
16	10	NO	0	9	12.3
17	10	YES	0	9	10.3
18	10	YES	3	9	10.4
19	10	NO	3	9	14.0
20	20	NO	0	9	20.3
21	20	NO	3	9	20.6
22	20	YES	3	9	20.0

line loss of 0dB is 20.3dB (line 9). The overall system noise figure for 6dB line loss only increases by 0.9dB. Thus, with high site noise, the line loss is less significant than where site noise is low.

Lines 12–15 illustrate the effect of a tower-top preamplifier at a site where the

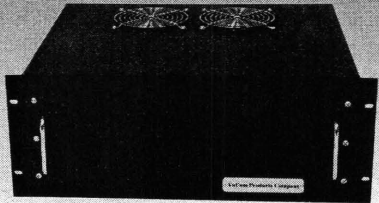
ambient noise level is at an absolute minimum. On line 12, no preamplifier is used, and the overall system noise figure is 9.5dB for a line loss of 0dB. On line 13, a preamplifier is used (line loss still 0dB), and the system noise figure is reduced to 4.2dB. This represents an improvement

of 5.3dB in the system noise figure—a significant improvement. On line 14, the line loss is increased to 3dB and with the preamplifier in place, the system noise figure only increases by 0.4dB. Thus, the preamplifier does an excellent job in improving the system performance in the presence of low site noise. Even with a 6dB line loss (line 15), the system noise figure is only 5.3dB.

Lines 16–19 show how a preamplifier affects system performance in moderate site noise levels. With a 10dB antenna noise figure and 0dB line loss, the system noise figure is 12.3dB without the preamplifier. With the preamplifier in line, the system noise figure is improved by only 2dB (line 17). If the line loss is increased by 3dB (line 18), the system noise figure only increases by 0.1dB.

Lines 20–22 show how system performance is affected by a preamplifier in a high site noise environment. On line 20, the preamplifier is not in line, and with 0dB line loss, the system noise figure is 20.3dB. If the line loss is increased to 3dB and still no preamplifier is in line, the system noise figure is 20.6dB. Line loss has little effect. On line 22, the preamplifier is inserted in line, and with 3dB line loss the system noise figure is 20.0dB. This represents an improvement of only 0.6dB compared to line 21, where the preamplifier was not in line. From this, it is obvious that a preamplifier is of little value at communication sites where the ambient noise level is quite high.

It is also apparent that in high ambient site noise, small line losses have little effect on the overall system performance. This would include insertion losses of cavity

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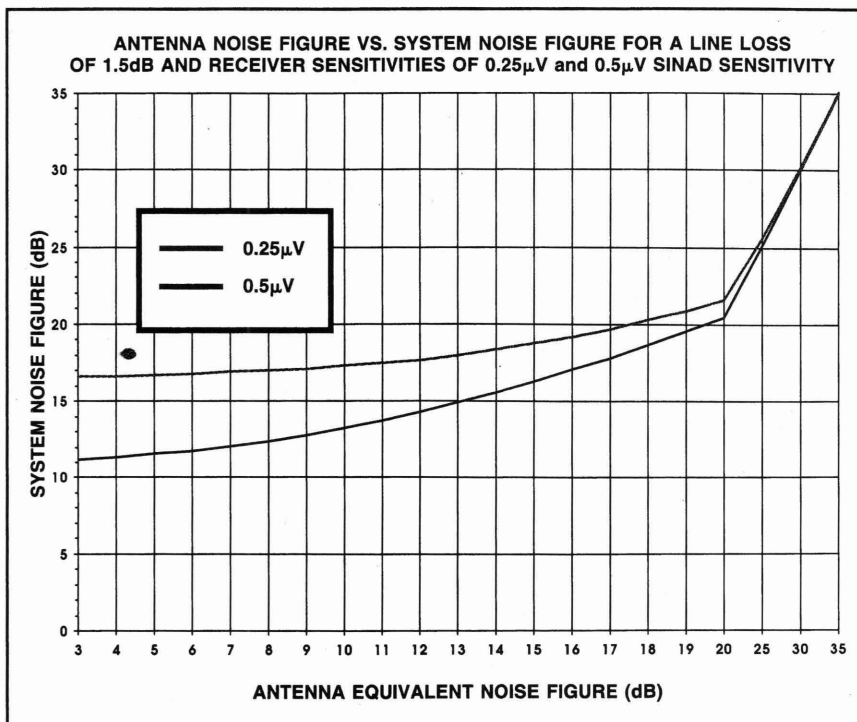


Figure 4. Plot of a receiver with a sensitivity of 0.25mV (red), compared to the plot of a receiver with a sensitivity of 0.5 μ V (blue).

filters and other insertion losses as well.

The isotee test

Suppose you perform an on-site test with the isotee to discover just how much site noise is degrading your receiving system. Suppose that you measure a degradation of 20dB with the antenna connected vs. the dummy load. Study the graph in Figure 4 above. This graph is for a receive system with a line loss of 1.5dB. The blue graph represents a receiver with a 12dB SINAD sensitivity of 0.5 μ V, and the red graph represents a receiver with a 12dB SINAD sensitivity of 0.25 μ V.

Now, if our isotee measurement were to indicate a degradation of 20dB with the antenna connected, the system noise figure would have been degraded by 20dB. Add the 20dB to the point of the vertical scale where the red graph intersects. Thus, 20dB plus 11dB = 31dB. Move up to 31dB on the vertical scale and over to the red graph and down to the horizontal scale, and we find the antenna has an equivalent noise figure of 31dB. If the external noise is reduced by 20dB so that the equivalent antenna noise figure is 11dB, then the overall system noise figure will increase less than 3dB. Move down the horizontal scale from 31dB to 11dB, up to the red graph and over to the vertical scale to about 13.5dB for the new system noise figure. Suppression of the noise much

below this level will yield little improvement.

Summary

Remember, site noise from several sources adds as the *root-sum-square* for non-coherent noise voltages. If a certain transmitter is degrading your system, it will do no good to suppress that source of noise much below the noise level of the general noise floor at the site. To be sure you are attacking the problem at the correct transmitter, observe the difference in noise degradation with the suspect transmitter up and down.

When the equivalent antenna noise figure is high, suppressing the noise level by approximately 6dB *plus* the amount of degradation will yield about the best improvement.

Further suppression will be of little benefit.

Until next time—*stay tuned!*

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Ott, Henry W., *Noise Reduction Techniques in Electronic Systems*, 2nd ed., John Wiley & Sons, 1988.



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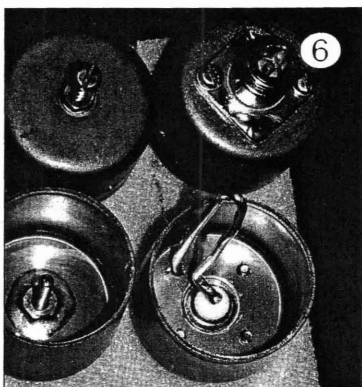
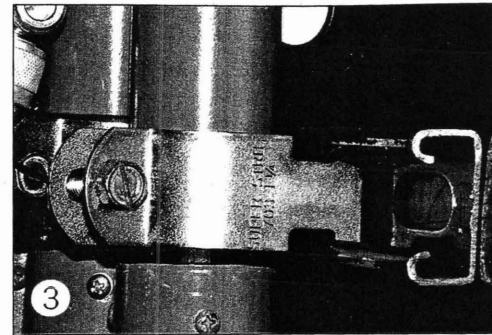
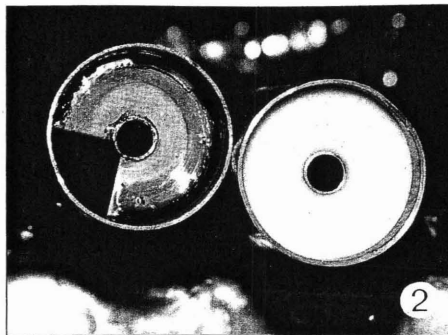
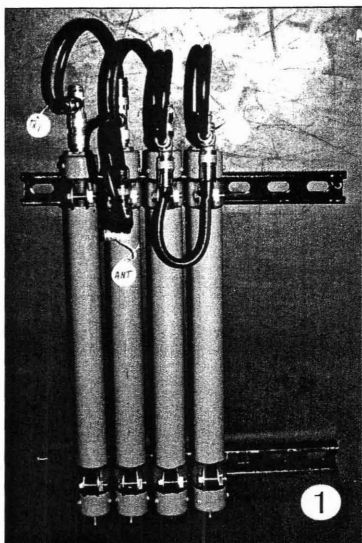
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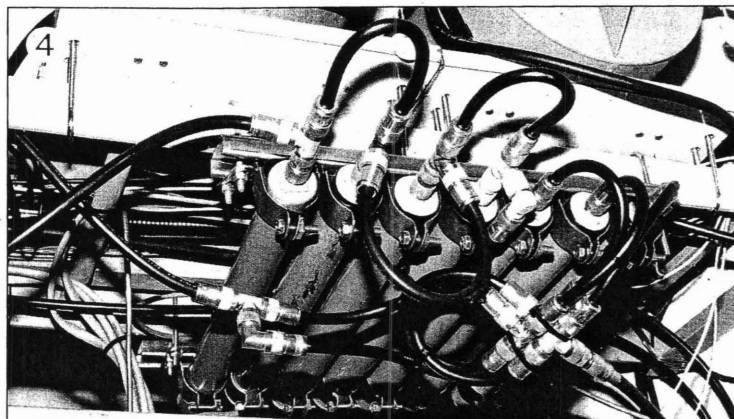


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Custom-fabricated VHF cavity filters. Clockwise: (1) a cavity duplexer arrangement; (2) link side and tuning side of cavity [center insulator for $\frac{3}{8}$ -inch pipe not installed]; (3) pipe strap attachment on strut channel; (4) three-way duplexer [three transceivers connected to one antenna]; (5) assembled cavity with black shrink tube; (6) both sides of pipe caps, link and tuning.



Make VHF cavities from over-the-counter hardware

Fabricating simple cavities may seem like work, but space, cost, flexibility and ease of tuning make the concept an attractive alternative.

By Patrick E. Buller

Often a situation arises where an expensive VHF cavity is not only big and bulky, but it also can result in more insertion loss than necessary. An economical approach is to fabricate smaller cavities from readily available materials. The experience can be rewarding and recreational.

The operation of most VHF-UHF cavities is based on the *quarterwave*

Buller is an electronics design engineer for the Washington State Patrol, Bellevue, WA. He is also a member of IEEE, NARTE, RCA, APCO, ARRL.

theory, where one end of a conductor is grounded, leaving the other end at a high-impedance (voltage) point. Minimum insertion loss is one of the goals for both *notch* and *pass* cavities. Notch and pass cavities differ only in the manner in which energy is inserted and removed. Pass cavities have two connections, one for the insertion of power, and the other to remove power at the tuned frequency. The pass cavity, when used as a notch cavity, will have one unused loop or port. A pass cavity can simultaneously operate as a notch cavity if its loop is tuned with a capacitor to a notch frequency and the pass cavity

is left tuned to the pass frequency. Notch cavities have one coaxial connection to remove energy at the tuned frequency. This discussion will be limited to the notch feature. Using large (4- to 10-inch) pass cavities for notch applications is an excellent approach, but it's overkill for wide frequency separation.

Q value

The term "Q" is a dimensionless value defining a "quality of merit" of tuned circuits. For a cavity, Q is a measurement of its narrow bandwidth of either pass or notch capabilities. A value of Q_u (unloaded Q) is directly

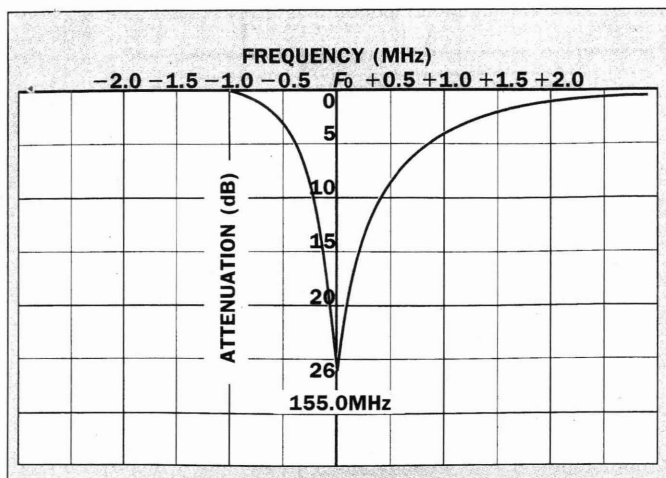


Figure 2. Decibel attenuation vs. frequency of a single-pipe cavity directly connected with a coaxial "T." The shape of the notch is that of a low-pass filter.

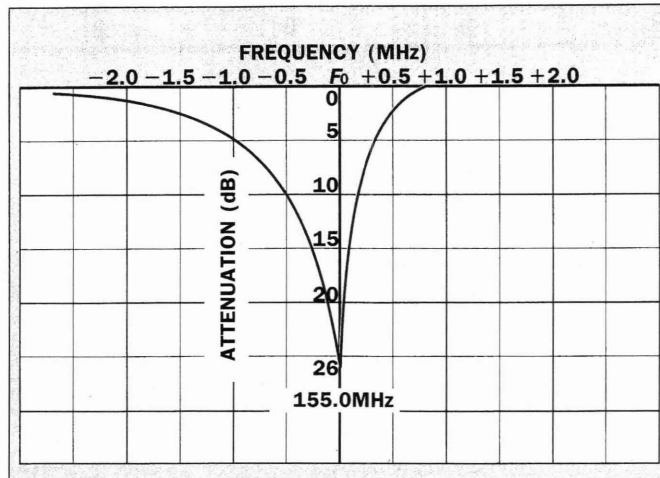


Figure 3. Decibel attenuation vs. frequency of a single-pipe cavity connected with a quarterwave stub, shifting the response to that of a high-pass cavity.

proportional to its volume. The larger the volume, the higher the Q. It is also influenced by the amount of coupling, called the "loaded Q," (Q_L). The relationship between selectivity (3dB bandwidth) and Q is demonstrated in the equation

$$Q = F_0 \div BW$$

where F_0 is the center frequency and BW is bandwidth.

As the Q is increased, the bandwidth decreases, or sharpens, the frequency

response. Increasing Q for more selectivity, the insertion loss increases, as determined by the equation

$$L = 20 \log_{10}(1 + (Q_U \div Q_L)).$$

For a bandpass cavity, Q_U and Q_L are transposed.

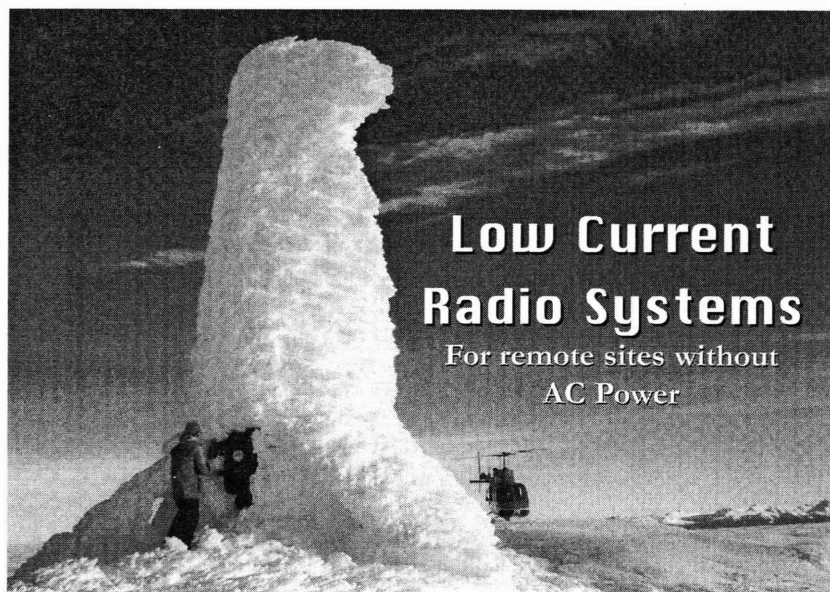
Values of Q range from 2 to 5 for broadband applications, such as a quarterwave coaxial cable, to that of a crystal having Q values beyond 10,000. The average Q for a 10-inch diameter

VHF cavity is about 350.

For coaxial cable use, cavities use *coupling loops*, usually identified by the insertion loss (i.e., 0.5dB, 1dB or 3 dB). The 0.5dB loop reduces the Q factor more than the 3dB loop. Another way of stating it is that a 0.5dB loop loads the cavity much heavier than the 3dB loop. Operating a cavity at $\frac{3}{4}$ -wavelength provides a higher value of Q, but it is generally not economical.

The assembly drawing shown in Figure 1 on pages 34 and 35 shows the design and construction of a cavity that has been in service for more than 15 years without failure. It was fabricated from 17.5 inches of 1.5-inch-diameter copper pipe. The only drawback is its temperature tolerance. The cavity is not recommended for applications where the ambient temperature will swing from -30°F to $+120^\circ\text{F}$. However, it is well-suited if the building can be maintained between 50°F and 90°F . If a reasonable temperature environment can be maintained, this may be just the thing for you. This design evolved because of the ease of repair and construction from readily available materials found in well-stocked hardware stores or at refrigeration-supply specialty stores. Coaxial connection to the cavity can be made with either type N or the common SO239 connectors. Test results show the "N" fittings are superior when three or more cavities are connected in any fashion.

The area of the loop (area between conductors) determines the amount of cavity coupling. Squeezing the loop conductors together decreases the total loop area, thereby reducing coupling, while increasing the loop conductor spacing increases coupling, which increases attenuation. The photographs of selected components on page 32 show the final form.



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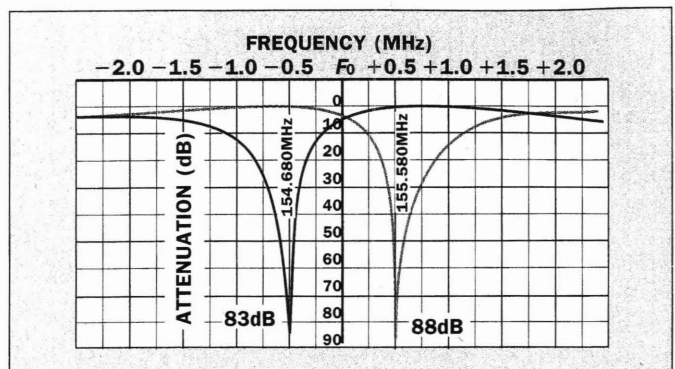
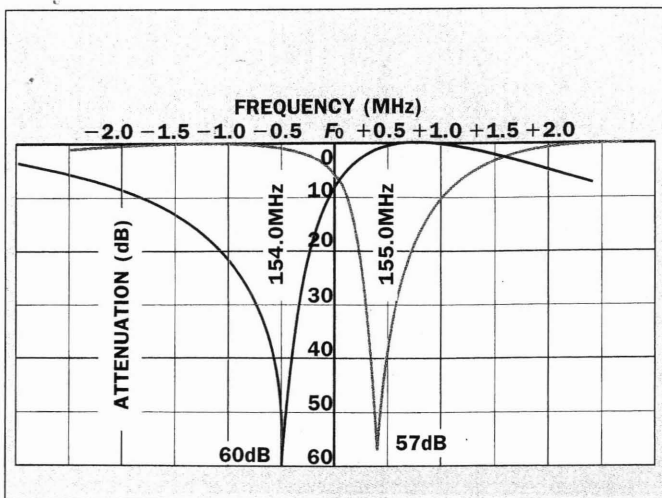


Figure 4. (Left) Decibel attenuation vs. frequency of a four-cavity assembly tuned at 1.0MHz spacing, with each cavity adjusted to 26dB of notch.

Figure 5. (Above) Attenuation vs. frequency of a six-cavity system at 900kHz spacing, with each cavity adjusted to 24dB of notch.

Initial adjustment

Each cavity should be individually adjusted to about 25dB if the frequency separation is 900kHz or more and to 20dB if approaching 800kHz spacing. Increasing the loop area will not only increase the notch attenuation but also the insertion loss close to the notch frequency. A spectrum analyzer with a tracking generator should be used as a tuning aid to display the notch while changes to the coupling are made.

Figure 2 on page 36 shows the frequency response of a single cavity

connected directly by a coaxial "T." Note that the shape of the notch is that of a low-pass notch filter. Adding a quarterwave coax from the "T" to the same cavity shifts the response to that of a high-pass cavity, as shown in Figure 3 on page 36. One-eighth-wave cable attached to the cavity makes the response uniform on either side of the resonance.

Three cavities interconnected with quarterwave cables will result in a nominal low-pass notch of 80dB. Three cavities, each having a quarterwave stub, and interconnected with quarterwave cables,

gives a high-pass notch of 85dB. Connecting each notch filter to a common point with another quarterwave cable yields a *cross-notch duplexer*. Figure 4 above shows the response of four cavities interconnected in this fashion and tuned at 1.0MHz spacing. In this case, each cavity was adjusted to 26dB of notch. Figure 5 above shows the response of a six-cavity arrangement, with each cavity adjusted to 24dB.

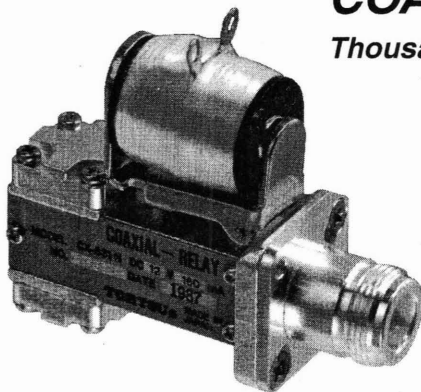
Filtering

One application for a notch cavity is the attenuation of an offending 155MHz transmitter generating noise on a receiver tuned 800kHz above or below the transmitted frequency. Adding a one-, two- or three-cavity assembly on the output of the transmitter reduces the offending noise with minimal attenuation to the transmitted signal. Where available space is a concern, this assembly can be mounted inside the base station cabinet. Installations are in service, using this design, that couple many receivers to one line with less than 0.25dB insertion loss. The common method of connecting two receivers, with an impedance-matching harness, nets about 3dB power loss. Another useful application is combining two 100W simplex base stations to one antenna. The spacing is 900kHz, and insertion loss is 1.25dB, resulting in 87W from each transmitter going to the antenna, as shown in Figure 5.

These cavities are most economical where the spacing between channels is greater than 800kHz. Anything closer than 800kHz with insertion loss less than 2.0dB requires a cavity Q greater than is possible with 1.5-inch-diameter pipe.

Mounting and cabling

The most economical method of mounting these cavities is to use the



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pipe-support systems developed for electrical conduit, often known by the trade names Unistrut or Superstrut. The pipe straps do not affect the tuning when fully tightened. Heat-shrink tubing of sufficient length is placed around the 1.5-inch copper tubing to isolate the copper metal from the steel mounting hardware. This prevents the dissimilar metal electrolysis from taking place. A benefit of the mounting scheme is that the entire coax and cavity is dc-isolated from the mounting assembly, which prevents unwanted ground loops.

Always use double-shielded or solid

outer-shield cable to interconnect cavities and also to connect to the equipment and antenna. Single-shield cable, such as RG8/U or RG58/U, has too much leakage and prevents maximum attenuation. It is also a source of extraneous radiation.

Success stories: Interference

The Washington State Patrol (WSP) was receiving severe interference on 155.580MHz from another transmitter on 154.740MHz located several hundred feet away in another public safety building. The WSP receiver desense

measured 35dB. A two-pipe cavity (40dB attenuation) was installed on the offending transmitter's transmission line, which eliminated the desense. The transmitting power output was 95W without the filter and 93W (0.09dB loss) after the filter was installed. Both parties are pleased with the results.

Another case involved the National Weather Service (NWS) transmitter at 162.550MHz, U.S. Customs at 165.2375MHz and WSP's repeater transmitter on 155.580MHz mixing, which created interference to WSP's repeater input. A four-pipe cavity was installed on the repeater input with two cavities tuned to the U.S. Customs output frequency and the other two cavities tuned to the NWS frequency, which eliminated the problem.

In the third case, the FBI needed a remote receiver at one of WSP's communications site. The frequency separation was more than 7MHz, making the installation of a two-pipe cavity more attractive than the installation of another antenna for the FBI. The two-pipe cavity coupled the two receivers to the single antenna with only 0.20dB of insertion loss. The common approach of using 75Ω quarterwave cables and a coaxial "T" would have resulted in more than 3dB of loss.

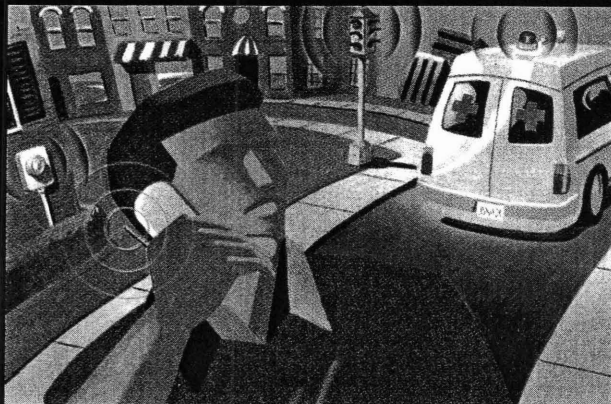
Combining radios

The most useful application is combining several transmitter and receivers to one antenna. Having separate antennas for each transmitter often generates problems in fringe areas. A mobile unit will find a hot spot for two-way communication on a given channel, but when the unit moves to another channel, communication is lost if the antenna is on the other side of the tower or, worse yet, if the antenna is much lower on the tower than the antenna used for the initial contact channel. Combining both radios to the same antenna provides identical communication capabilities. In other words, the hot spot is the same for both channels. An added benefit to this application is fewer antennas required on the tower. As many as four transceivers can be coupled on one antenna with these pipe cavities.

Maintenance

If the insulator inside the 3/8-inch pipe fails because of a voltage breakdown, it can easily be replaced. Remove both end caps and insert a rod inside the 3/8-inch pipe to force the Teflon insulator out. Insert a new insulator from the tuning side of the cavity.

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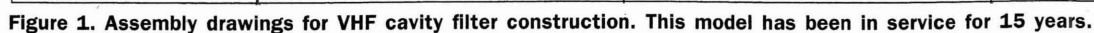
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Using SINAD voting combiners in mobile radio networks

SINAD voting systems improve wide-area coverage on common channels. Voting frees operators from selecting zones and channels. Diversity reception can improve performance.

By A.O.J. Woodfield

Traditional land mobile radio systems include communications among a dispatcher at a fixed location and operators who use various mobile and portable units.

The base station might be installed at the fixed location where dispatching takes place. If that location is unfavorable for radio coverage, the communications range may be limited.

Woodfield is manager of Uni-Lab New Zealand, Christchurch, New Zealand. When this article was written, he was systems manager at Tait Communications, Christchurch.

The base station might be installed at a more favorable location and remotely controlled from the dispatch location to improve coverage.

The system may use a repeater station on a high building or geographic feature to relay communications throughout a fairly large area.

When even greater range is required, two or more repeaters may be connected by radio links, microwave links or telephone lines to combine their coverage areas. Linking allows all members of a fleet to monitor and use a common audio channel. Such systems usually require the mobile users to select the appropriate channel for the area in which

they are located. Because the correct channel may not be immediately obvious to users traveling across many different repeater coverage areas, it is not uncommon to find many incidents of lost calls or poor communications because users select the wrong channel.¹

Mobile receiver scanning is one way to solve the problem. The scanning receiver switches automatically from channel to channel until a signal is found. Unfortunately, a simple scanner locks onto any received transmission, not necessarily on the best available channel.

Another solution is to use quasi-synchronous transmission methods in which all transmitters, controlled by highly stable oscillators, broadcast simultaneously on a common channel.² Quasi-synchronous transmission is expensive, and its maintenance requires great care and relatively expensive test equipment.

Mobile voting

A mobile voting system includes receiver scanning, but with an important addition: When a signal is received on any of the scanned channels, the mobile scans once more through all of the other channels and measures the received signal strength on each one. It selects the channel with the strongest signal.

The process requires a fast mute response time in the mobile and a fast signal strength measuring capability. A typical mobile scanning five channels may require 450ms (milliseconds) to complete the cycle before receiver audio is switched on. It takes about 50ms per channel to unmute; another 30ms per channel to measure the signal

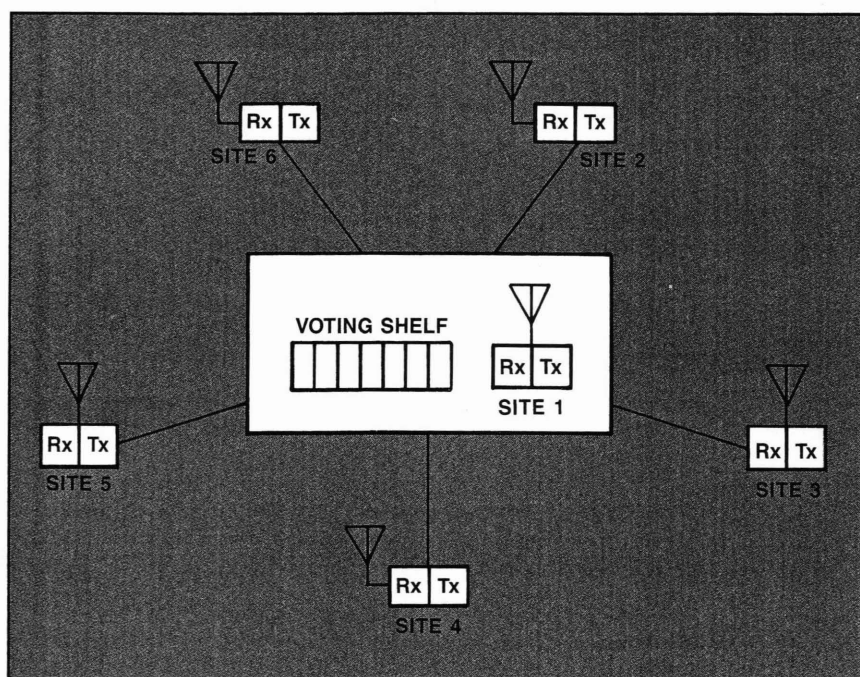


Figure 1. In the more common star network, satellite receivers are placed throughout the coverage area. They feed signals to a centralized voting shelf.

strength; and another 50ms to compute the result, select the channel and unmute again to allow the operator to hear the message. Until then, the audio output remains silent.

When continuous-tone controlled squelch system (CTCSS) is involved, the process takes more time. If the detection of a valid tone is left until the last stage in the process, it may increase the

delay by 100ms to 200ms for a 650ms delay. For a system with CTCSS and 10 channels, a second may elapse before audio is heard, a delay that is unacceptable to most operators.

Some mobile voting systems require the permanent presence of transmitter carriers to operate properly. Such a requirement is inefficient, and it creates a potential source of interference.

Another problem with these wide-area coverage systems is the rapid rise in channel requirements. In many cases, the greater channel numbers can result in significantly reduced channel loadings and a consequential rejection of channel requests by local regulatory authorities.

The increasing use of portable transceivers adds problems for the network designer. Although a hand-held receiver's sensitivity may approach that of the usual mobile receiver, the lower transmitter output level, typically 5W (+37dBm), coupled with the lower efficiency of a shortened, helical antenna, results in significantly reduced portable coverage. To cope, system designers may add fill-in repeaters with additional linking, adding to the channel and linking requirements.

For larger networks, link channel requirements can be substantial. If UHF radio links are used, basic equipment design limitations, such as audio signal-to-noise performance and distortion figures, can limit the number of possible links.

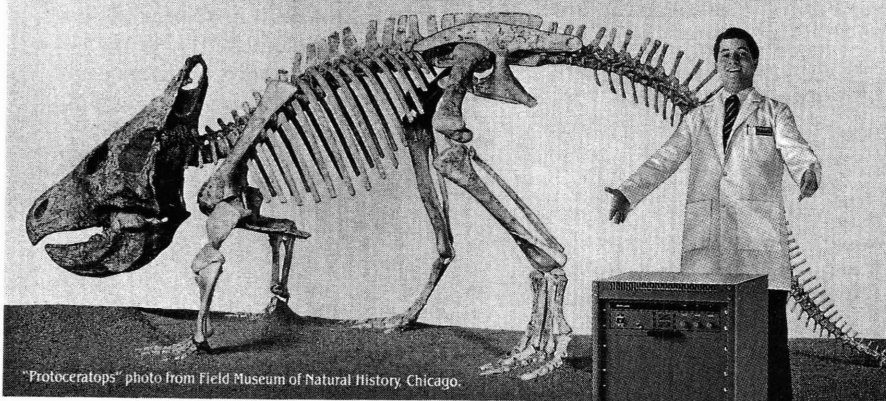
SINAD voting

SINAD voting systems offer a solution. They help to cover a wide area with multiple radio channels; reduce RF channel and linking requirements; and support selective signaling, data transmission and CTCSS in much the same way they are used in conventional repeaters.

The SINAD voting system continuously measures received audio from receivers within the coverage area. The SINAD of all of these audio signals, usually represented by a fluctuating dc voltage, is passed to a comparator. The comparator analyzes all valid incoming signals. The best signal is selected and fed to the transmitter or transmitters covering the desired area.

The best channel selection may be made either on a continuous—almost syllable-by-syllable—basis, or on a once-per-transmission basis. The once-per-transmission mode may be desirable for sites that cover low-signal-level areas where communication with mobiles otherwise may be lost when received SINAD levels are consistently low. This mode may be desirable, moreover, when data transmission is involved

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because frequent phase changes in received signals would be undesirable.

Network configurations

Two basic SINAD voting system configurations are available to the system designer, the star and the linear-spur.

► *Star network*—Figure 1 on page 30 shows the more common star network, in which satellite receivers are placed

throughout the coverage area. They feed signals to a centralized voting shelf. This central site may include the transmitter for the coverage area and another receiver, but this is not essential.

The links may be wireline or radio, and they may use tone-on-idle or E/M signaling to indicate receiver gate operation.

Selected audio may be passed to a

control room or to the local transmitter for repeater applications. If the system transmitters are located around the coverage area, the voted audio can be selectively routed to them or passed to all transmitters. A word of caution: If multiple transmitters are used simultaneously on identical frequencies and if their coverages overlap, quasi-synchronous simulcasting must be used to avoid problems associated with frequency drifts and offset.³

► *Linear-spur*—Figure 2 on page 36 shows the linear-spur network. It resembles the star network and is more appropriate for covering long-and-thin-shaped areas, such as highways and railways. The figure shows coverage along a highway.

Remote base stations feed their signals to a central control room via three voting systems operating in series. With this scheme, any mobile can access the system anywhere along the highway using a common, single mobile transmitter frequency, and all mobiles use a common receive frequency. The voters automatically retransmit on the correct transmitter or on a transmitter specifically selected by the central control room operators.

SINAD voter operation

On the test bench, SINAD measurements normally are simple to make. A 1kHz test tone modulates the RF test generator, and the receiver's audio output level is measured two ways: directly and through a sharp, 1kHz notch filter to give the SINAD value. Nonetheless, when only normal speech or signaling is present, measuring SINAD is more difficult.

One method that represents the various schemes available in commercial voting systems is shown in Figure 3 on page 38. Each incoming received audio signal is amplified logarithmically, increasing low-level signals and providing less gain for high-level signals. This stage is followed by an envelope detector that detects the gain-limited audio delivered by the previous stage.

The valley detector then measures the difference between the detected audio signal's peak, corresponding to the SIGNAL + NOISE + DISTORTION component of a SINAD measurement system and the signal's valleys, which correspond to the

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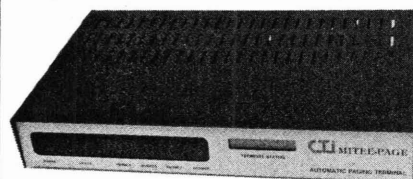
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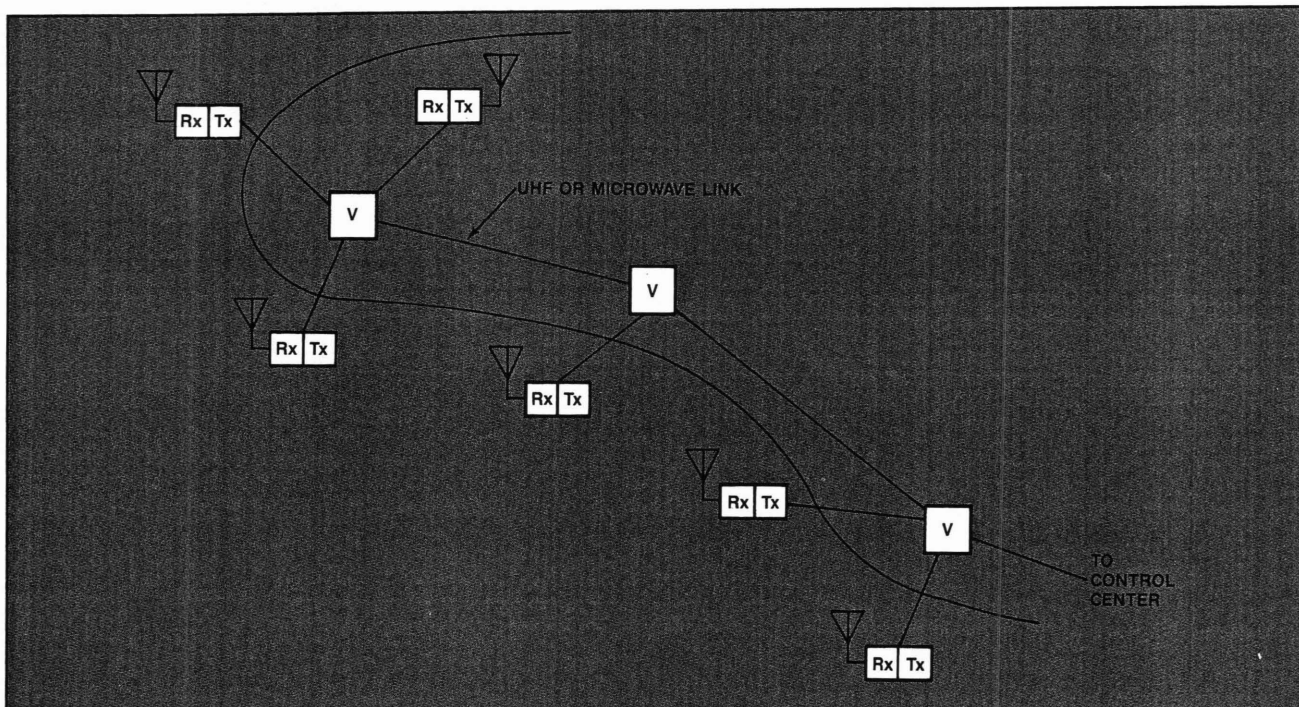


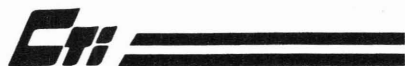
Figure 2. The linear-spur network resembles the star network. It is more appropriate for covering long-and-thin-shaped areas. In this example, a highway system's remote base stations feed signals to a central control room via three voting systems operating in series.

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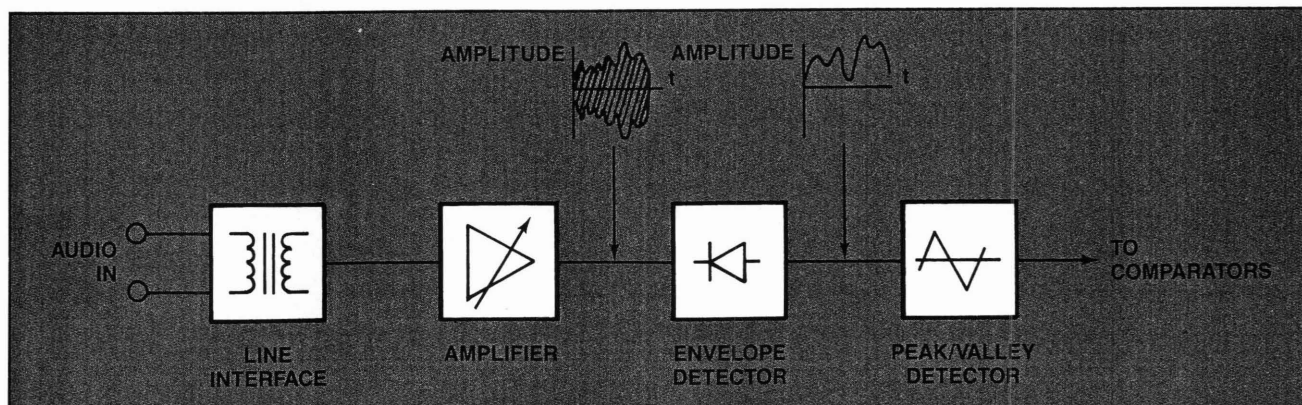


Figure 3. In this configuration of a commercially available voting system, incoming received audio signals are amplified logarithmically. An envelope detector detects gain-limited audio delivered by the previous stage. The valley detector measures the difference between the detected audio signal's peak, corresponding to the SIGNAL + NOISE + DISTORTION component of a SINAD measurement system and the signal's valleys, which correspond to the no-signal NOISE + DISTORTION of SINAD. The resulting dc output of this difference-detector then represents the instantaneous received SINAD.

no-signal NOISE + DISTORTION of SINAD. The resulting dc output of this difference-detector represents the instantaneous received SINAD.

This brief overview ignores some highly critical areas of the various circuit blocks, especially the important dy-

namic range constraints, the required system signal timing and the decision logic that surrounds these stages to ensure that only valid signals are analyzed and passed to the comparators.

It is essential that no audio compression or limiting occurs in the receiver

audio stages or intermediate linking equipment prior to being presented to the voter for analysis. If compression or limiting occurs, then the log amplifier may limit or amplify the low-level signals excessively. The following envelope detector stage may then limit, and the

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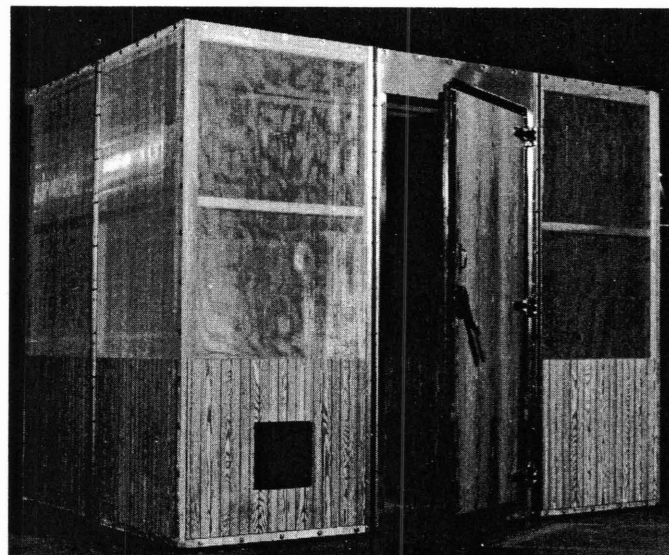
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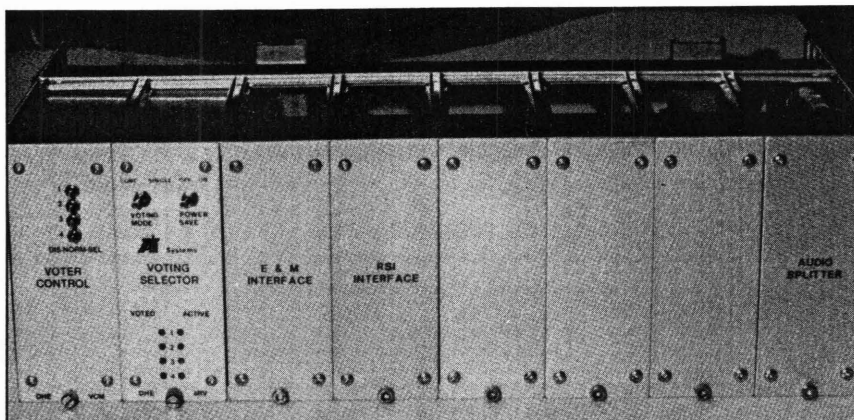
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A UHF diversity rack shows two monitored UHF shelves (top) and a voting system for diversity combining of the two UHF receivers on the UHF shelves.

resulting response of the valley detector may indicate a poor SINAD for that channel. If the channel is limiting badly, this situation inadvertently may indicate the truth—that a bad-quality signal is being received—but for the wrong reason.

Other SINAD measurement techniques adopted in SINAD voter equipment attempt to avoid the problem through the

use of window detection and other methods, but all systems can be affected badly by gain limiting and compression.

Time constraints

Fast voting system operation is crucial to successful radio system operation, especially if selective calling or data transmissions are to be used. Excessive delays in decision logic require

the user's mobiles to adopt lengthy lead-in delays prior to sending selective calling (selcal) or data transmissions. In the latter case, the lead-in delay can lead to extremely poor overall system performance, even with the best high-speed data terminal equipment. If the system must wait for 500ms before the transmission channel is ready, for example, a five-digit selcal sequence at 20ms per tone, which usually takes 100ms to 150ms to send, will require 600ms to 650ms to achieve the same result.

Decision logic

Decision logic must ensure that only valid received signals enable the SINAD voter detectors and faults such as lost or disfunctional audio lines are able to be cleared automatically by detection circuits and timers, or manually, if required. For example, the detectors almost invariably see valid channels that have no audio signals as having the best SINAD because noise and distortion on faulty lines is extremely low. This problem can be avoided with tone-on-idle

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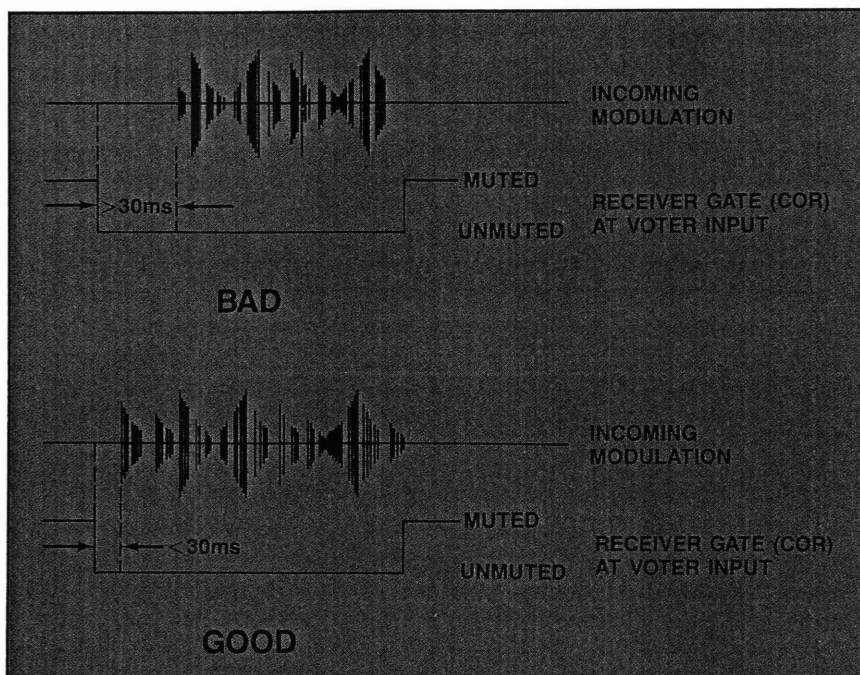


Figure 4. If selective calling or automatic number identification (ANI) signals are sent at the beginning of transmissions, the selcal and ANI data will be partially or completely missed if erroneous timing occurs. The drawing shows examples of good and bad gate signal sequences.

signaling timeout timers, hard-wired E+M lines or similar methods.

Timing

It is essential to ensure that signaling applied to the voter is timed correctly. In particular, a receiver gate signal must not arrive before valid audio arrives. If it does, the voter automatically selects the channel with the receiver gate signal and remains on it until audio arrives at the inputs. Then it makes a proper selection.

If the voter time constraints provide hysteresis to hold a selected channel for a nominal 200ms (normally, an acceptable period), it may take 250ms to select a correct channel following an invalid selection based on the early arrival of a receiver gate signal.

If selective calling or automatic number identification (ANI) signals are used in such a system, and if they are sent at the beginning of transmissions, then the selcal and ANI data will be missed partially or completely if erroneous timing occurs. Figure 4 on the left shows

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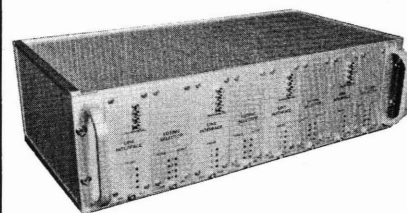
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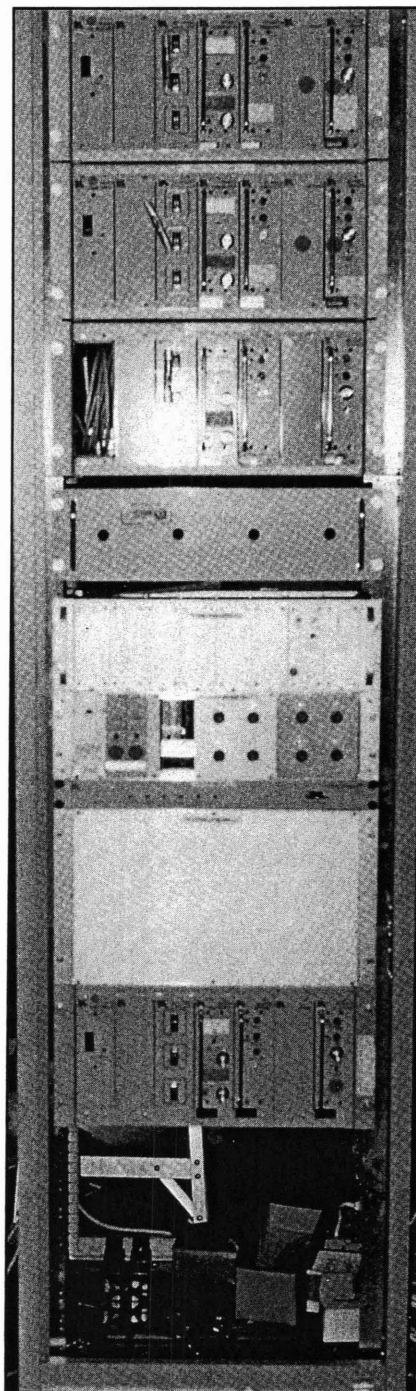
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A typical four-channel SINAD voting system shelf with an optional four-wire E/M wire interface, received SINAD indicator (RSI) interface and an audio splitter for audio retransmission.

good and bad examples of gate signal sequences.

The use of CTCSS signaling within voting systems is one example in which such timing sequences can be experienced, and careful design checks must

be made to ensure correct operation. System alignment must be handled carefully to ensure that all factors are checked.

Frequency response

Where several voting systems are set up in series, as in the linear-spur configuration, the system frequency response is composed of the sum of the individual equipment frequency responses. If voters are in sequence, and if each has a 300Hz-to-3,000Hz, ± 3 dB frequency response characteristic, the output from a series of three such systems will exhibit a 9dB rolloff at 300Hz and 3,000Hz. This is not the way to engineer a professional sound on a mobile network.

It is important to keep edge rolloff and in-band ripple to a minimum. In networks designed by the writer, the specification adopted is 300Hz to 3,000Hz ± 0.2 dB. A simplex frequency response tailoring network is added to each voter to ensure that the specification is met. With five such voters in series, the network output should be better than ± 1 dB across the frequency range.

System alignment

The following describes a typical alignment sequence for a voting network. It is essentially valid for almost all system configurations.

(1) All outstation radio equipment should be aligned and tested fully, and correct line levels should be set. If outstations use CTCSS tones, they should be checked, and the modulation levels should be measured.

(2) Align and test radio linking equipment, especially where CTCSS regeneration is to occur. Where fixed lines are to be used, they should be tested to ensure that they are noise-free and that their frequency responses are correct.

(3) Align and test the voting shelf, initially using a SINAD voter alignment test set if available and using valid system audio line levels.

Do not use continuous audio tones for SINAD voter testing and alignment because continuous tones do not provide the audio detectors with the necessary peak-and-valley sequences. SINAD voter test sets provide gated tones that partially simulate speech or signaling. The

gated tones allow correct alignment and permit operation to be tested fully. A proper test set lets the voter be tested completely and independently of the networks and allows network problems to be isolated more rapidly.

(4) Check the SINAD voter shelf for at least 3dB hysteresis between channels to ensure that minimal interchannel selection "chatter" occurs in service

when mobiles present nearly equivalent SINAD inputs to multiple receivers. Hysteresis of more than 6dB should be avoided to ensure that smooth interchannel selection occurs within the voter. Signal acquisition time should be less than 50ms after receipt of a valid gate signal, and channel hysteresis should provide a hold time of about 200ms. Shorter hold times produce excessive

"chatter," and longer times delay valid channel selections.

(5) Check audio output levels from the voter and the presence of a valid transmitter key signal when a valid receiver gate input is received.

(6) Test the signal/gate timing between each outstation and the inputs to the voter shelf. Use a modulated RF signal generator as a source for the receiver, whose RF output is gated slowly on and off (at two-second intervals, for example), and observe (using a storage os-

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Shorter hold times produce excessive "chatter," and longer times delay valid channel selections.

cilloscope at the voter) the arrival of the receiver gate signal and the received modulation. Modulation may precede slightly or be coincidental with the gate signal, but it must not be more than 20ms or 30ms behind the arrival of a valid receiver gate signal. Check that no modulation gaps occur during the switching between receivers on a common shelf.

(7) Check the audio level at each transmitter site to ensure correct modulation levels are maintained. If at all in doubt, check the link or line audio frequency response to ensure an optimum audio response is present—preferably better than ± 1 dB at the band edges for any individual audio path.

Other voter system uses

► **Diversity**—SINAD voting systems can be used for receiver diversity applications. Diversity techniques improve the communications path between mobiles and the base stations by combining multiple signals and reducing the periodic deep fading that occurs when mobiles move about the coverage area.

Space diversity improves the received

(continued on page 78)

(continued from page 46)

signal by combining the outputs of multiple receivers (usually two) whose antennas are spaced by more than a quarter wavelength. Polarization diversity uses a similar setup with two antennas of different polarization. The technique works because signals received by separated antennas are unlikely to suffer

simultaneous fading.⁴

A voting system can be used to select automatically the best output signal from two or more receivers fed from diversity antenna systems. Typically, average received signal strengths are approximately 4dB better for two-receiver diversity combining and as much as 8dB better when four receivers are combined, as compared to a single receiver.

The greatest improvement is in the reduction in deep, sharp fades that may reach 10dB or more, even when two receivers are combined with either space or polarization diversity.⁵

► **Improved coverage**—Where well-defined areas must be covered, such as on a highway running north and south from the radio base station site, using two receivers connected to individual yagi antennas pointed north and south, for example, provides well-defined, high-gain coverage. This method uses standard antennas, as compared to the alternative of purchasing a specially designed, phased-antenna system.

► **Duplicated systems**—By combining two receivers through a SINAD voter, system redundancy for improved reliability can be provided. Redundancy may be especially useful at remote sites and sites with difficult access.

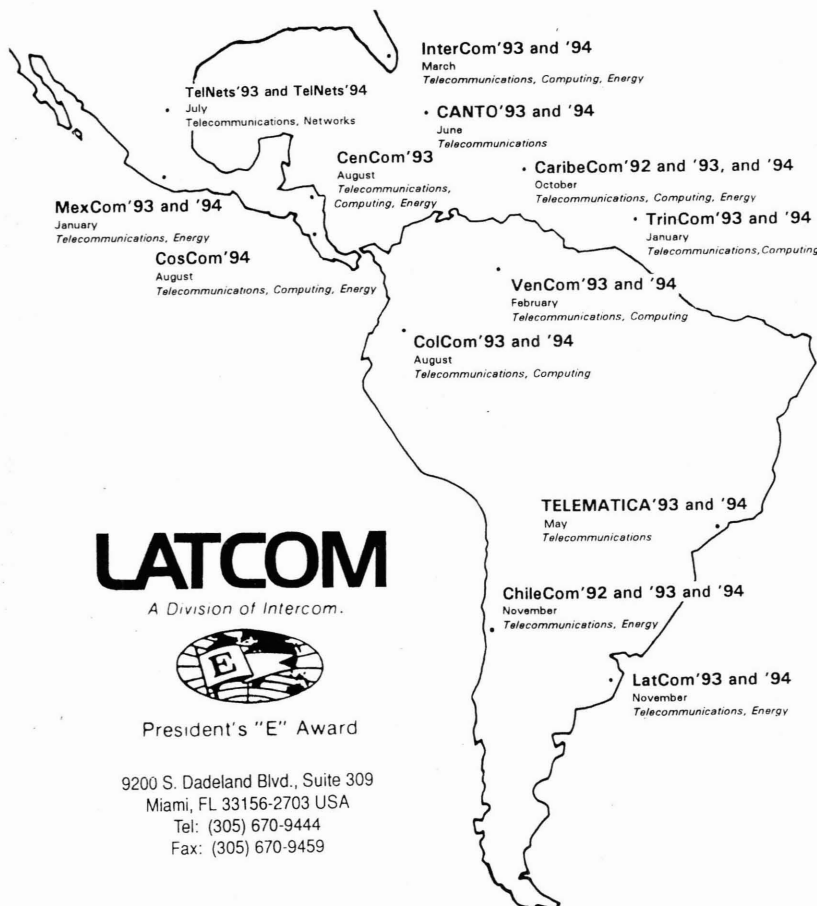
SINAD voting systems provide a powerful tool for planning systems that require wide-area coverage on common channels. SINAD voters can provide a means for co-located receivers to be added into a communications network, particularly where polarization-, space- or directional-diversity reception systems are being considered.

Care is required when implementing such systems, but experience has shown that SINAD voting systems cover large areas with ease. User satisfaction is especially high when operators no longer have to select zones and channels.

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Every toolbox needs one of these

The RF sampler is an important test accessory for any field technician

Before retirement, I was actively engaged in field service, which often required RF tests and measurement procedures that, in turn, required special equipment, test devices and accessories. The smaller RF test devices and accessories were kept in what I called my "RF Toolbox." Without a simple — but special — accessory, it might be impossible to run a specific RF test on equipment. Therefore, this special toolbox was always kept close at hand.

One of the most important test accessories to have in your toolbox is the RF sampler. Many years ago I learned how to fabricate a simple RF sampling device from a UHF Tee connector. The idea was to remove the center pin from the connector, using long-nose pliers, and cut the pin to a short length. A hacksaw was used to file a screw slot into the end of the pin so it could be reinserted into the connector using a regular flat-blade screwdriver.

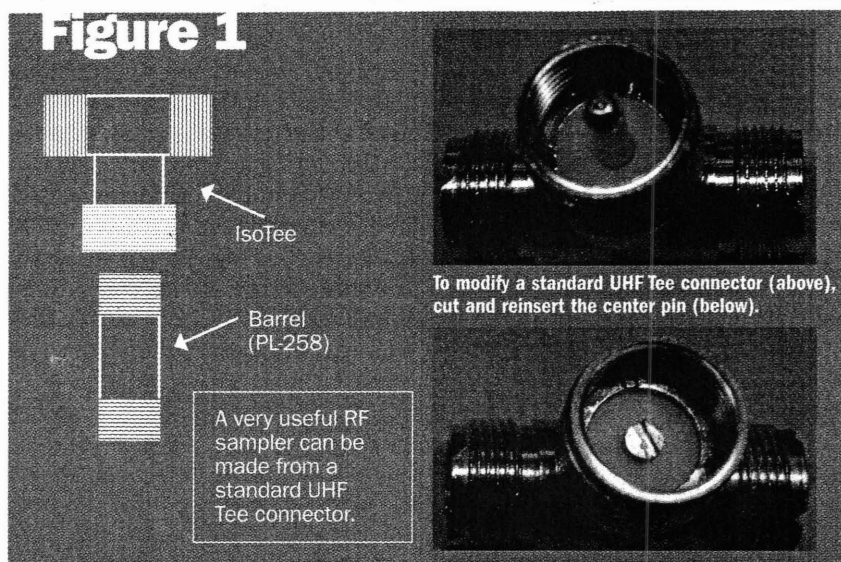
When reinserted into the Tee connector, the end of the pin would be just about flush with the insulator. This homemade RF sampler is usually referred to as an "IsoTee." Figure 1 shows a modified connector that has been connected to a barrel to create an IsoTee. The photos show a regular Tee connector before modification and one that has been modified. (I don't know who can truly claim credit for the idea, but that person has my thanks.)

It is a good idea to mark the IsoTee somehow so that you can easily distinguish it from a regular Tee connector, as confusing the two in a test setup could result in disaster. In this regard, I have seen some technicians insert a #8 screw into the end of the barrel to increase the coupling through the arrangement. In any case, be sure that there is no direct or DC contact through the IsoTee. If there is, the purpose is defeated, and equipment damage might result.

Figure 2 shows how the IsoTee is used as a signal sampler to tap into a line running significant RF power. This provides a small sampling of the signal to operate the frequency counter or other sensitive RF instrument. The isolation provided by the IsoTee pre-

is connected. The signal generator is adjusted to again produce 12-dB SINAD at the receiver output. The new signal generator level then is recorded.

The difference between the required signal generator levels (in decibels) is equal to the amount of noise degrada-



vents overloading the frequency counter input, and because of the isolation, there is no noticeable impedance bump on the transmission line.

Not only is the IsoTee useful for sampling RF power on a line, but it also is useful for injecting a signal into a line for receiver testing. One of the primary uses for the IsoTee is the effective site sensitivity test, which determines how much degradation the local site noise was causing to the receiver sensitivity. Bench sensitivity means very little in the presence of high site noise.

The basic setup for running this test is shown in Figure 3 on page 30. First, the 50-ohm termination is connected to the left side of the IsoTee and the signal generator is set to produce 12-dB SINAD at the receiver output. The output level of the signal generator is measured in dBm and recorded. Then, the 50-ohm termination is disconnected and the antenna

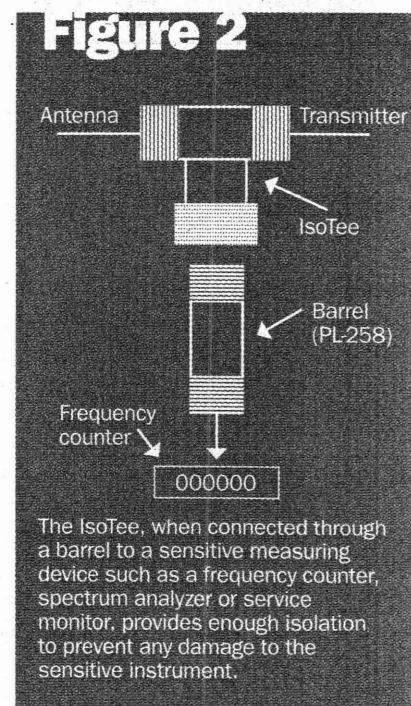
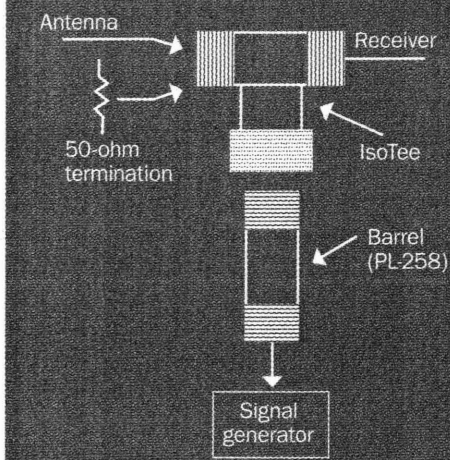


Figure 3



To determine how much receiver sensitivity is degraded by local site noise, the 50-ohm termination is connected to the left side of the IsoTee and the signal generator is set to produce 12-dB SINAD at the receiver output. The signal generator level is recorded in dBm. Then the antenna line is connected and the signal generator level is adjusted to again produce 12-dB SINAD at the receiver output. The difference between this signal generator level and the previous level is equal to the amount of degradation being caused by local site noise.

tion experienced at the site. Due to the isolation, high signal generator output levels will be required to produce 12-dB SINAD at the receiver output. This simple test setup works quite well because we are not measuring the absolute signal level. We are only concerned with the difference between signal levels, in decibels, when the termination is connected

versus when the antenna is connected.

Typically, a signal generator should be terminated into a 50-ohm load. It is standard practice to use something like a 6-dB pad between the signal generator and the receiver. At higher frequencies the barrel and the standard Tee connector can cause an impedance mismatch. For the die-hard purist, the following

setup might be used. Figure 4 shows how a pad between the generator and the Tee connector and a 50-ohm termination on the connector can reduce any impedance mismatch problem caused by the IsoTee setup. If the return loss without the pad is 10 dB and a 6-dB pad is used, the return loss with the pad will be 22 dB. A return loss of 10 dB is equal to a standing wave ratio (SWR) of around 1.92:1, and a return loss of 22 dB is about equal to an SWR of 1.17:1.

The reason for this dramatic improvement with a pad is that we get a two-for-one improvement in return loss. Because the signal has to travel through the pad in the forward direction and back through the pad in the reverse direction, the signal is attenuated twice. In other words, a 6-dB pad gives us an additional return loss of 12 dB. Good engineering practice calls for the use of pads in many test procedures.

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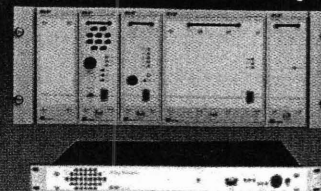
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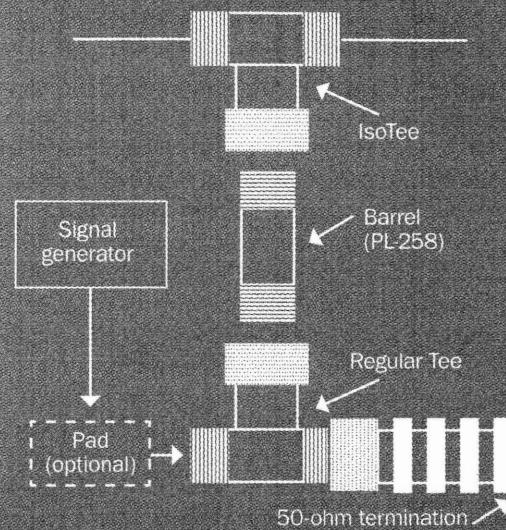
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struct their own, RF Industries makes a commercially available RF sampler. The RFA4059A sampler can be used with many different types of RF connectors that are available from the company. Connectors supplied with this RF sampler are N-male on one side and N-female on the other side. This sampler can be used with any of the Unidapt connectors available from RF Industries. The company also offers a kit of adapters for the RFA4059A RF sampler that makes it convenient to connect almost anything to anything. The sample port of the RF sampler is through the BNC connector, and the degree of coupling is adjustable.

Bird Electronic Corp. also offers a commercially available RF sampler that can be mounted to the Bird 43 wattmeter. The amount of coupling is adjustable. Signal samplers of various types are available from most of the wattmeter manufacturers. These

Figure 4

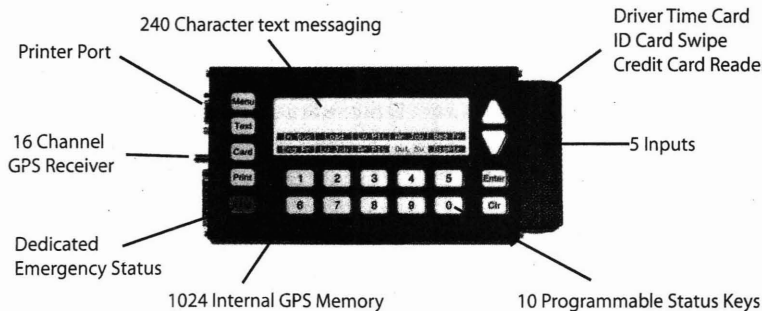


This test setup will present a proper termination to the signal generator. It is common practice to pad the output of a signal generator. Typically, a 6-dB pad is used at the signal generator output. Compared to an open-circuited output at the signal generator, a 50-ohm load will reduce the voltage from the signal generator by 6 dB. If a 6-dB pad is used, the signal level is reduced by another 6 dB. This results in a 12-dB reduction compared to the setup in Figure 3. However, the signal generator should still have enough reserve to overcome the loss of the IsoTee and the 12-dB loss.

include directional and non-directional samplers that are built into wattmeter "slugs," as well as sample ports that are mounted on the wattmeter housing.

RF samplers can be classified as directional or non-directional, adjust-

able or fixed, frequency-compensated or non-compensated. It is impossible to describe all of them here. Choose the types that are best-suited to your needs and keep them handy in the RF toolbox. Until next time, *stay tuned!* ■



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Good to know

The relationship between field intensity and signal level should be understood by all RF technicians

By Harold Kinley

The term "field intensity" is used to indicate the intensity level of the signal at a given distance from the radiating antenna. A simplistic definition of field intensity might be: the level of the signal "in the air." The basic unit of measure of field intensity is volts per meter, or V/m. In practice, millivolts per meter (mV/m) or microvolts per meter ($\mu\text{V/m}$) are used to indicate typical field intensity levels.

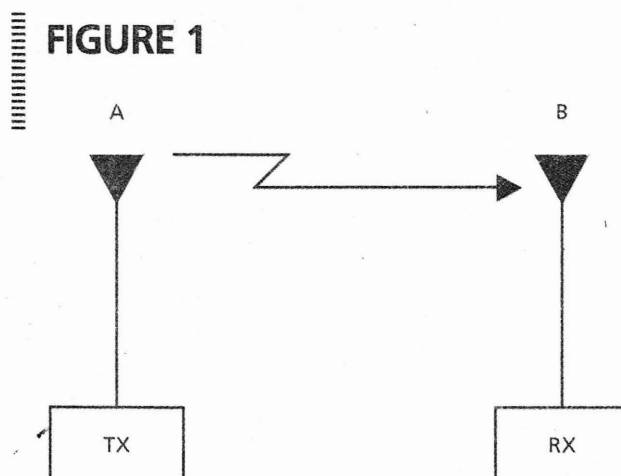
On the other hand, the signal level across the antenna terminals—or at the input/output of a receiver or stage of a system—usually is specified in terms of volts (V), millivolts (mV) or microvolts (μV), or decibels referenced to 1 milliwatt (dBm) or 1 watt (dBW). Actually, the terms dBm and dBW are used to represent power

levels but easily can be converted to voltage as long as the system impedance is taken into account. For example, most RF technicians and engineers know from memory that 1 microvolt (1 μV) is equivalent to -107 dBm in a 50-ohm system. This article will help you to correlate the field intensity at the antenna to the signal level at the input to the receiver in practical situations.

Figure 1 shows a transmitter connected to a half-wave dipole antenna and a receiver connected to a half-wave dipole antenna. The transmitter power is 1 kilowatt, or 1,000 watts. For this example, assume that the line losses are zero. The distance between the two antennas is 1 mile. The formula for calculating the field intensity in volts per meter, shown in Equation 1, assumes free-space propagation.

Since frequency is not part of the formula, we must assume that the calculated field intensity is independent of frequency. Plugging the numbers from the example above into Equation 1 yields a field intensity of 0.1375 V/m (137.5 mV/m) at point B, the receiving antenna.

This does not mean that an RF signal level of 137.5 mV appears across the antenna terminals (with a 50-ohm load connected). This is where frequency becomes a factor. The difference between the field intensity level at the antenna and the voltage appearing across the antenna terminals differs, in decibels, by the antenna factor. The antenna factor (for half-wave dipoles) is defined by the formula in Equation 2.



The effective radiated power from antenna A is 1,000 watts. The distance between antenna A and antenna B is 1 mile (1,610 meters). The transmission lines on each end are lossless. The field intensity at B is determined by the formula in Equation 1.

If we plug a frequency of 160 MHz into Equation 2 we find that the antenna factor is 12 dB/m. Recall that the field intensity at antenna B in Figure 1 is 137.5 mV/m. The signal level appearing across the antenna terminals and at the receiver input (assuming no line loss) is equal to the field intensity (e) minus the antenna factor (k), or 137.5 mV reduced by 12 dB. In terms of voltage, 12 dB represents a reduction equal to 25% of the original voltage, or $137.5/4 = 34.4$ mV appearing across the antenna terminals and the input to the receiver. The formula for the signal level appearing across the antenna terminals is shown in Equation 3.

The antenna correction factor can be negative or positive. For a half-wave dipole, an antenna correction factor of 0 dB/m occurs at a frequency of approximately 40 MHz. Above 40 MHz, the correction factor is a positive value and below 40 MHz the correction factor is a negative value. For a given field intensity, a negative antenna factor will produce a higher signal level across the antenna terminals than a positive antenna factor. This is because a negative value substituted for k in Equation 3 produces a smaller denominator than would a positive value.

Field intensity often is expressed in terms of decibels referenced to 1 microvolt per meter and is written as dBu. That is, 0 dBu is 1 μ V/m and 6 dBu is 2 μ V/m. When the field intensity is stated in dBu it is easier to it convert to a signal level across

EQUATION 1

$$e = \frac{7\sqrt{P}}{d}$$

where e is field intensity in volts per meter,
P is effective radiated power in watts and
d is the distance between antennas in meters.

EQUATION 2

$$k = 20\log\left(\frac{2\pi F}{300}\right) + 1.5$$

where k is antenna factor in decibels per meter (dB/m)
and F is frequency in MHz.

EQUATION 3

$$s = \frac{e}{10^{\left(\frac{k}{20}\right)}}$$

where s is the signal level in the same units as e and
k is antenna factor in dB/m.

EQUATION 4

$$s = e - k$$

where s is signal level in dB μ V (decibels referenced to 1 μ V) across the antenna terminals,
e is field intensity in dBu and
k is antenna factor in dB/m.



the antenna terminals — this is done by simply adding or subtracting the antenna factor in dB/m. The formula in Equation 4 can be used for this purpose.

For example, if the field intensity at the antenna is stated as 12 dBu and the antenna factor is stated as 12 dB/m, the signal level in dB μ V across the antenna terminals is 12 dBu – 12 dB/m = 0 dB μ V, or simply 1 μ V. Don't confuse the use of the formulas in Equations 3 and 4. Equation 3 is used when the field intensity is in volts/meter or a related unit, such as mV/m or μ V/m. Equation 4 is used when the field intensity is stated in dBu.

A simple way to convert between field intensity and signal level across the antenna terminals is just to remember that for a half-wave dipole the correction factor at 40 MHz is approximately 0 dB/m. Above 40 MHz, the half-wave dipole antenna correction factor is positive and below 40 MHz the antenna factor is negative. A positive correction factor means a lower signal level across the antenna terminals, while

TABLE 1

MULTIPLIER	Δ dB (POWER)	Δ dB (VOLTAGE)
$\times 1.25$	+1	+2
$\times 1.60$	+2	+4
$\times 2.0$	+3	+6
$\times 10$	+10	+20
$\times 1/1.25$ (0.8)	-1	-2
$\times 1/1.60$ (0.625)	-2	-4
$\times 1/2.0$ (0.5)	-3	-6
$\times 1/10$ (0.1)	-10	-20

a negative correction factor means a higher signal level across the antenna terminals, for a given field intensity. For each octave (doubling of frequency) the antenna factor increases by 6 dB. Therefore, at 80 MHz the antenna factor for a half-wave dipole is 6 dB/m. If the field intensity is 20 dBu, the signal level across the antenna terminals is 20 – 6 = 14 dB μ V, or 14 decibels above 1 microvolt, or approximately 5 μ V. At 20 MHz, a given field intensity would produce a signal level across the half-wave dipole antenna terminals that is 2 times (6 dB greater than) that at 40 MHz.

Consider an antenna with a field intensity of 50 μ V/m. The frequency is 160 MHz and the antenna is a simple half-wave dipole (0 dBd gain). Since this is 2 octaves above 40 MHz, the antenna factor is 6 + 6 = 12 decibels — a positive value. This means that the signal level across the antenna terminals is 12 decibels below the



A simple way to convert between field intensity and signal level across the antenna terminals is just to remember that for a half-wave dipole the correction factor at 40 MHz is approximately 0 dB/m."

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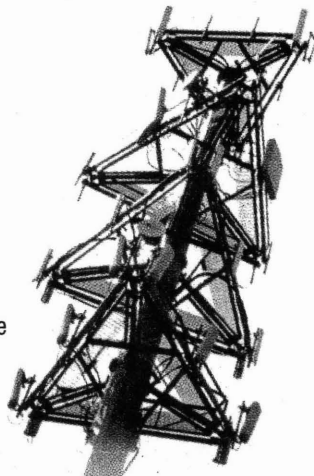
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Antennas used in pagers and handheld units may be quite lossy compared with a half-wave dipole and therefore produce much less signal voltage or power at the receiver input."

field intensity. Thus, the signal level at the antenna terminals is $50/4 = 12.5 \mu\text{V}$. Now suppose that the antenna in Figure 2 has a gain of 6 dB referenced to a half-wave dipole, or a gain of 6 dBd. This would increase the signal level across the antenna terminals by 6 dB to 25 μV .

Further suppose that the antenna is connected to the receiver through a coaxial cable that has a loss of 2 dB. Assuming a signal level of 25 μV at the antenna terminals, what is the signal level at the receiver input? Table 1 provides a handy reference for making that determination. The gray section of the table correlates multiplication factors with equivalent decibel increases in power or voltage. For example, the first figure in the table indicates that a 25% increase in the power level equals an approximate increase of 1 dB. A 25% increase in the voltage level represents a 2 dB increase.

Meanwhile, the blue section of the table shows that the reciprocal of the multipliers in the gray section of the table will produce the same change in decibel level, but in the



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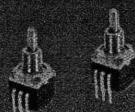
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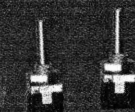
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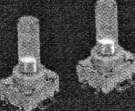
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opposite direction. According to Table 1, a 2 dB signal voltage loss in the transmission line would translate to a multiplication factor of 0.8. Thus, a 25 μ V signal level multiplied by 0.8 is equal to 20 μ V at the input to the receiver.

Another convenient number to remember is that 1 microvolt equals -107 dBm (in a 50-ohm system). Let's convert the 25 μ V signal level at the antenna terminals to dBm by using the figures in Table 1. The ratio of the signal voltage to 1 microvolt is 25:1. Consulting Table 1 we find that we can get to 25 by using the multipliers of 10 (20 dB), 2 (6 dB) and 1.25 (2 dB)— $10 \times 2 \times 1.25 = 25$. Adding the corresponding decibel figures, we get $20 + 6 + 2 = 28$ dB. Thus, 25 μ V is 28 dB above 1 μ V. Now, simply add (algebraically) 28 dB and -107 dBm to get -79 dBm. This is the signal power at the antenna terminals. This is worth repeating: The term dBm represents signal power but easily can be converted to voltage as long as we know the system impedance. The signal power at the receiver input is reduced by the line loss of 2 dB to result in a signal level of -81 dBm.

When using the formula in Equation 2, remember that this applies only to a half-wave dipole (a gain of 0 dBd). Antennas that exhibit gain over a half-wave dipole will produce more signal voltage across the antenna terminals for a given

field intensity level. Some antennas actually exhibit a loss compared to a half-wave dipole. For example, antennas used in pagers and handheld units may be quite lossy compared with a half-wave dipole and therefore produce much less signal voltage or power at the receiver input. ■

Harold Kinley is the author of three books, including Standard Radio Communications Manual and The Radioman's Manual of RF Devices. A certified electronics technician, he holds an FCC First Class radio telephone license and an amateur radio operator Extra class license.

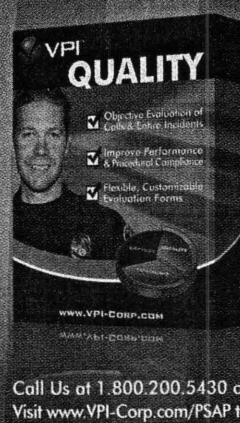
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Single-point grounding for communications sites

Single-point grounding is the most critical element of a three-part process involving effective bonding and grounding, transient voltage surge suppression and structural lightning protection.

By Bruce A. Kaiser

"Single-point grounding" is a current buzzword in protecting communications sites from the ravages of lightning and general ground transients.

Kaiser is president of Lightning Master, Clearwater, FL.

What is single-point grounding? Actually, the term, as used here, is a misnomer. Single-point ground electrical potential referencing has nothing to do with the grounding system itself or with the provision of only a single path from a particular piece of equipment back to the internal grounding provisions. The grounding system is determined by the soil conditions and other factors at a specific site. As a practical matter, it is impossible to

provide modern equipment with one, and only one, ground return path to the internal grounding provisions.

Single-point ground electrical potential referencing means connecting all site equipment to the grounding system at a single point, or, more precisely, bonding all communications site equipment to the grounding system in such a way that all of the equipment samples ground potential at one, and only one,

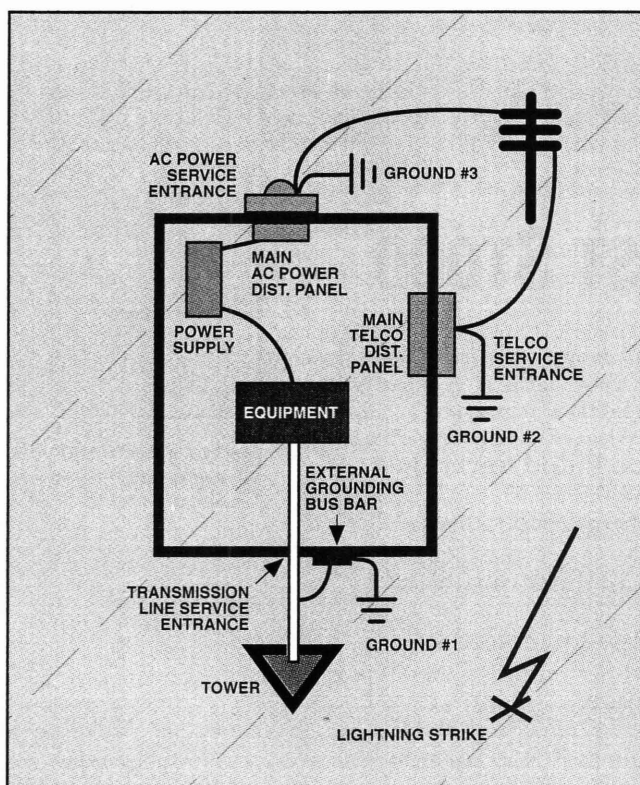


Figure 1. At many sites, the ac power service entrance is at one end of the equipment shelter, the telco service entrance is at one side, and the transmission line service entrance is at the end opposite the ac power entrance. Each service is grounded where it enters the shelter. When lightning strikes near the site, the ground potential changes across the site due to secondary effect.

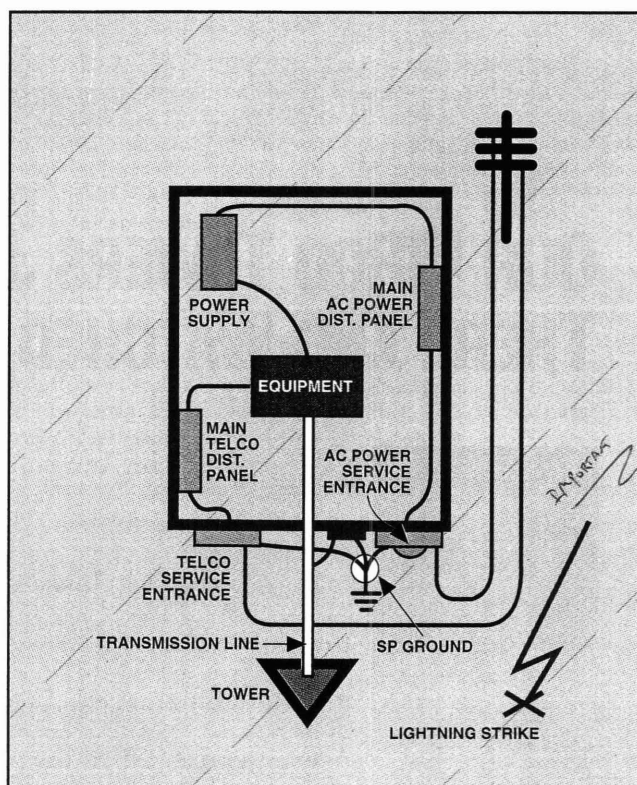


Figure 2. Make the transmission line potential the site reference potential by placing a large copper bus bar on the equipment shelter's outside wall below the transmission line service entrance. Install a transmission line grounding kit and bond it to the bus bar. Run a large lead from the bus bar to the site grounding system to place the RF side of the site equipment at the potential of the bus bar.

point (hence, single-point).

'Potential' damage

A change in electrical potential in and of itself does not damage equipment. It is a difference in electrical potential across or within a piece of equipment that causes damage. If one radio is connected to both the coaxial cables from the tower and to a power supply, and there is a difference in electrical potential between those two

services, that difference in potential equalizes within the radio, causing damage or accelerated wear. The same risk applies to two pieces of equipment communicating with one another through data lines. If there is a difference in electrical potential between the two pieces of equipment, that potential equalizes through the data lines within one or both pieces of equipment. (Communications site equipment, in this case, refers only to electrical or electronic

equipment. Communications site equipment does not include door frames, air conditioning ducting or the tower.)

Single-point ground definition

Envision an imaginary plane at or just below the floor of the site. All of the site equipment is appropriately bonded together above this plane, and an appropriate grounding system is established below this plane. Those two systems are bonded together at one, and only one, hole through that plane. Therefore, all equipment within the site is at the ground potential of that single point. The single-point ground is defined as the point at which the unified ground passes through this plane.

Several ground potentials are involved in a typical communications site. The first set of ground potentials is associated with ac power, telephone lines and RF transmission lines to the site. The second set is associated with the various electrical and electronic equipment chassis. There are other sources, too, but the following information covers these two.

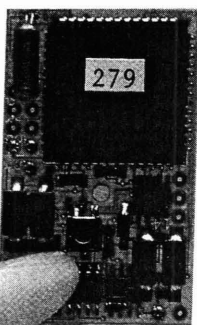
Changes in ground potential across a site may be caused by a variety of factors, but the most dramatic and familiar is lightning. During a thunderstorm, the charge at the base of the storm cloud induces a shadow of opposite charge on the surface of the earth beneath it. As the storm cloud blows along through the atmosphere, this ground charge, or more accurately, earth surface charge, is dragged along the surface of the earth beneath it.

Cloud-to-ground lightning results when the difference in electrical potential between the storm cloud charge and the ground charge exceeds the dielectric of the intervening air. That air breaks down in a series of steps, and the lightning strike occurs. When lightning strikes a particular point on earth, the ground charge at that point is vacated relative to the surrounding ground charge. The surrounding charge rushes to the point of the strike.

Ground charge distribution

Looking at the distribution of ground charge immediately after a strike, it is apparent that the ground charge potential will change with distance from the strike. A point close to the strike will be at a quite different potential compared to another point at a greater distance from the strike. If a ground rod is driven at each of these points and the electrical potential between the two rods could be measured during a lightning strike, the measurement would indicate a dramatic and almost instantaneous difference in potential between the ground rods. If a power supply were grounded

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to one of the rods and a radio powered by it were grounded to the other, during a strike the radio would see a difference in electrical potential between its power input and chassis. This difference in electrical potential will equalize within the radio.

At many sites, the ac power service entrance is at one end of the equipment shelter, the telco service entrance is at one side, and the transmission line service

entrance is at the end opposite the ac power entrance. (See Figure 1 on page 10.) Each service is grounded where it enters the shelter. When lightning strikes near the site, the ground potential changes across the site due to secondary effect. The propagation rate of the ground charge causes a difference in ground potential (dashed lines) at each service entrance ground. The equipment within the shelter therefore sees a difference in potential between the RF

input at the potential of ground No. 1, the power input at the potential of ground No. 3, and the telco input at the potential of ground No. 2. That difference in potential is equalized across and within the equipment, causing wear or damage.

Transmission line ground

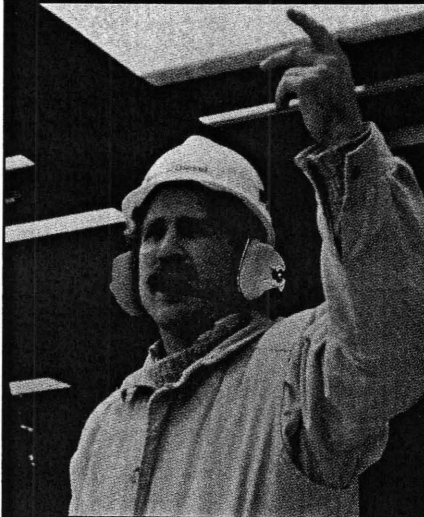
With respect to the ground potential of the services to the site, the most difficult service potential to control is the transmission line from the tower. This potential is difficult to control in part because the tower usually is several yards away from the shelter, and the tower's potential is constantly changing.

The transmission line ground potential can be limited relative to the rest of the site, or the transmission line potential can be made the reference potential for the remainder of the site equipment. Based on experience with Florida cell sites, the second solution is easier to implement—and actually more effective.

Grounding configuration

The transmission line potential can be made the site reference potential (single

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*When lightning strikes
near the site, the ground
potential changes across
the site due to
secondary effect.*

point) by placing a large copper bus bar on the equipment shelter's outside wall just below the transmission line service entrance. A grounding kit is installed on the transmission line (in addition to the grounding kits already installed at the top and bottom of the tower) and bonded to the bus bar. A large lead is run from the bus bar to the site grounding system. This configuration places the RF side of the site equipment at the potential of this bus bar. (See Figure 2 on page 10.)

Then we route the ac service to the site so it meets the equipment shelter just to one side of the transmission line service entrance. At the main disconnect, the ac neutral and ground are bonded together, and a ground lead extends from that box. This ground lead should be attached to either the transmission line service entrance bus bar or to the large ground lead from that bus bar above the point at which it attaches to the grounding system. This assures that the ac power side (and low

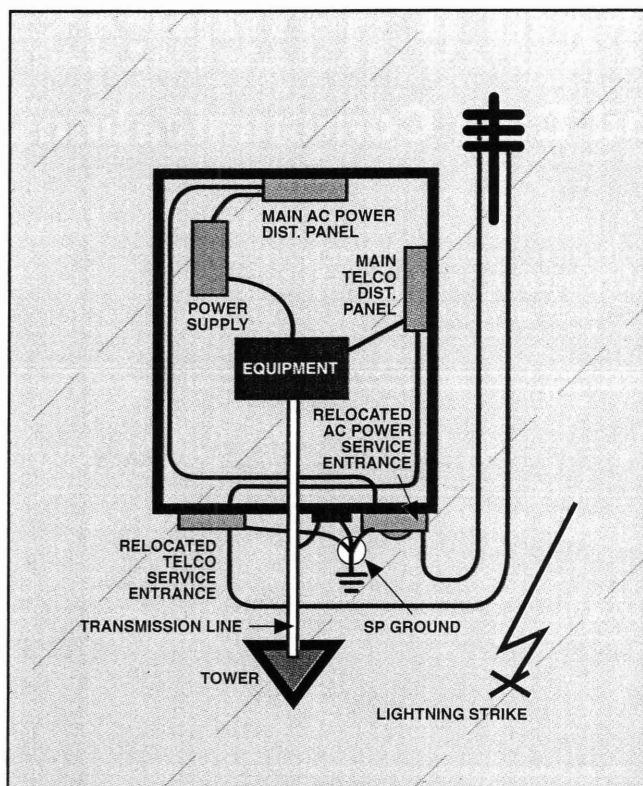


Figure 3. One of two things can be done about transmission line ground potential: It can be limited relative to the rest of the site, or the transmission line potential can be made the reference potential for the remainder of the site equipment. Based on experience with Florida cell sites, the second solution is easier to implement and actually more effective, as shown above.

voltage power side) ground of the site equipment will be at the same potential as the transmission line bus bar and therefore

at the same potential as the RF side of the equipment. (See Figure 3 above left.)

Then the telco service is routed to the

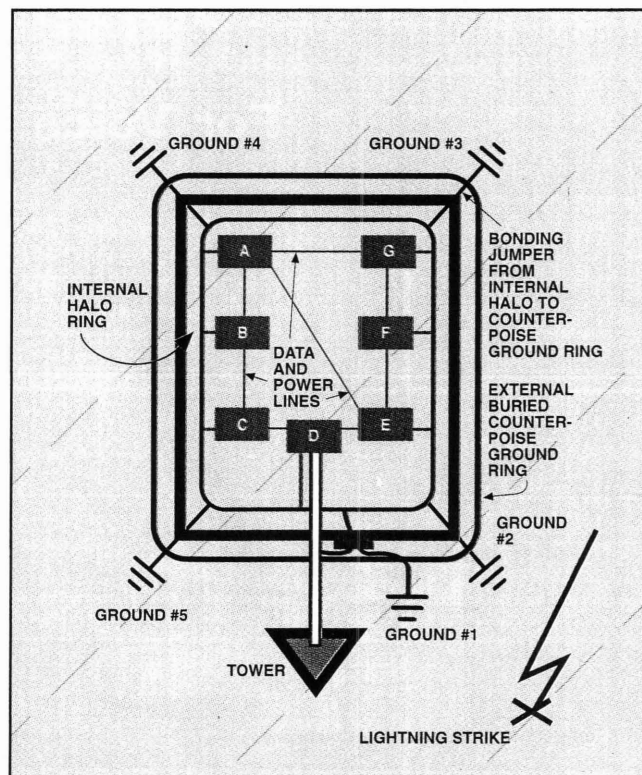


Figure 4. In many sites, an internal halo ground is installed at the shelter perimeter and bonded to ground at all four corners. Site equipment is bonded to the halo at a convenient location, so any given piece of equipment samples the closest grounding point. During a nearby lightning strike, the ground potential at one corner of the site could differ vastly from the ground potential at any other corner.

site so it meets the equipment shelter on the opposite side of the transmission line service entrance. The ground lead extending from the telco service box is attached either to the transmission service entrance bus bar or to the large ground lead from that bus bar above the point where it attaches to the grounding system. This configuration assures that the telco side of the site equipment is at the same potential as the transmission line bus bar and therefore at the same potential as the RF and power sides of the equipment.

With this arrangement, any difference in ground potential among services to the site is equalized at the single point, and the site equipment samples all ground potentials at one, and only one, point: the single-point ground. A nearby lightning strike would produce the same step potential across the site, but because all site equipment is attached to services sampling the ground potential at only one point (and there is no other ground reference potential), the equipment sees no current flow across or within it to equalize electrical potential.

Converting existing sites

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to design into a new site. But what about an existing site? The internal distribution routing of ac power and telco lines is already established, and it is difficult and expensive to change it. The answer is to leave the existing internal distribution routing intact and just move the service entrances. With the ac power, move the main disconnect to a location near the transmission line service entrance and ground it to the single point. Then run a

"primary jumper" internally back to the existing distribution panel. This way, none of the distribution routing needs to be changed.

The same technique should be employed with the telco service. Move the demarcation or service entrance to the transmission line service entrance and establish its ground at the single point.

As expensive as it may be to move service entrances, it is less expensive than re-

placing damaged equipment or equipment that wears out prematurely. At least it is not necessary to reroute the internal distribution of the ac power and telco lines within the shelter.

Chassis grounding

The next area of grounding concern is chassis grounding. In many sites, particularly older ones, an internal halo ground was installed around the perimeter of the shelter near the ceiling or floor and bonded to ground at all four corners of the shelter. (See Figure 4 on page 16.) Site equipment then was bonded to the halo at the closest or most convenient location, so that piece of equipment sampled ground potential at the closest grounding point. During a nearby lightning strike, the ground potential at one corner of the site could be

During a nearby lightning strike, the ground potential at one corner of the site could be vastly different from the ground potential at any other corner.

vastly different from the ground potential at any other corner. The chassis potential of each piece of equipment will be closest to the potential of the ground at the nearest corner of the building. The chassis potential of component A, as shown in the figure, will be close to the potential of ground No. 4. The chassis potential of component C will be close to the potential of ground No. 5, and the chassis potential of component E will be close to the potential of ground No. 2. Because many pieces of equipment communicate with one another through data lines, and all of them share the power source, electrical potential differences equalize within and through the site equipment.

Interior bus bar

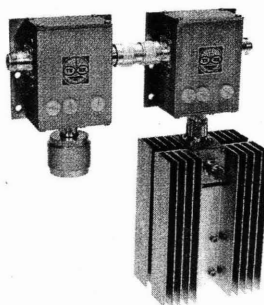
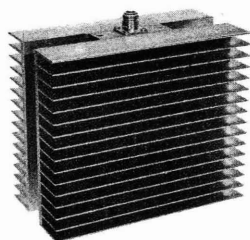
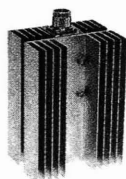
The solution is to bond the chassis of each piece of equipment to ground at the same point. (See Figure 5 on page 22.) The easiest method is to install a copper bus bar inside the shelter on the opposite side of the same wall as the external transmission line bus bar. Each piece of site equipment is bonded to the internal bus bar, preferably with a bonding wire run from

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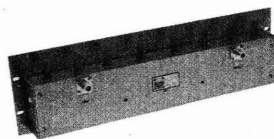
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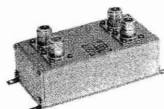
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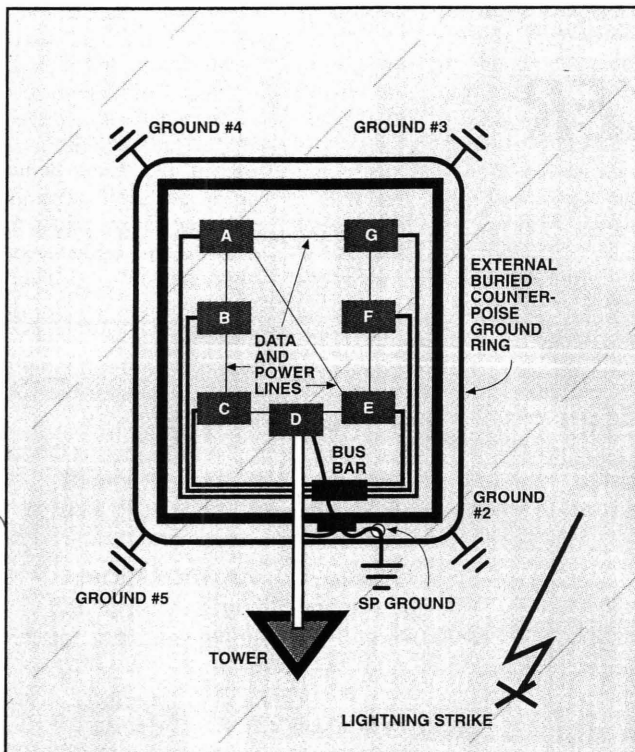


Figure 5. The solution to equalize electrical potential differences within and through the site equipment is to bond the chassis of each piece of equipment to ground at the same point. The easiest method is to install a copper bus bar inside the shelter on the opposite side of the same wall as the external transmission line bus bar and bond each piece of site equipment to the internal bus bar.

each chassis directly to the bus bar. The internal bus bar is bonded through the wall to the external bus bar. Ideally, equipment ground leads should be arranged across

two internal bus bars according to how they produce or absorb surges, and whether they require non-isolated or isolated signal ground references. Whatever the case, all

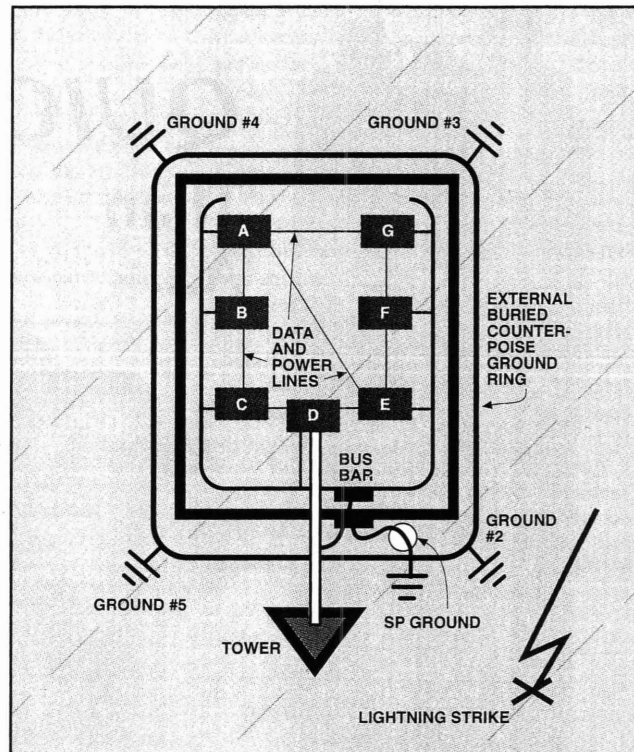


Figure 6. What about existing sites with internal halos grounded at multiple points to the external buried counterpoise? Ground the internal bus to the external bus ground lead. Cut the internal halo at the end of the building away from the transmission line service entrance. Cut each wire grounding the internal halo to the external buried counterpoise and remove them from the walls.

equipment should be chassis-grounded to one potential.

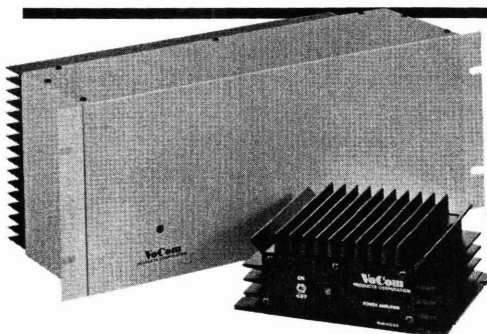
A nearby lightning strike now would produce the same step potential across the site, but because all site equipment chassis sample ground potential at only one point, and there is no other ground reference potential, the equipment sees no current flow across or within it to equalize electrical potential.

Buried counterpoise

The real purpose of the external buried counterpoise around the shelter is to protect people and the site from the adverse effects of step potential across the site during a nearby lightning strike; therefore, the counterpoise is important and should be part of any site grounding scheme. Even so, its purpose is not to provide a convenient "bus" to which equipment grounds should be tied.

The single-point chassis ground is effective and easy to design into a new site. But what about existing sites with internal halos grounded at multiple points to the external buried counterpoise? Fortunately, many problems at an older site can be solved with a good pair of wire cutters. First, ground the internal bus to the ex-

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ternal bus ground lead. Then cut the internal halo at the end of the building away from the transmission line service entrance, eliminating the large ground loop. Then cut all wires grounding the internal halo to the external buried counterpoise and remove them from the walls. This modification leaves each equipment chassis grounded only at the single-point bus bar. During this process, be certain not to leave any piece of equipment ungrounded

inadvertently. (See Figure 6 on page 22.) One of the ironies in this case is that it is less expensive and uses less material to ground the site properly than to install a multiple-point ground.

Direct lightning strike

In all of the previous examples, a nearby lightning strike has been the cause of damage. A direct strike to the tower, equipment shelter or utility service pole on the

site produces an even more impressive change in ground potential across the site. With a properly designed and installed single-point grounding system, the site should survive the ground potential change of even a direct strike. The electromagnetic pulse (EMP) and other effects of a close strike, and methods of mitigating their damage, are another story, but they are not covered in this article.

A few matters become apparent when considering single-point grounding of all services and equipment. The most obvious is that it is much easier to design a new site to this standard than it is to modify an existing site. Equipment shelters can be specified and ordered from the manufacturer with all service entrances at the same end of the shelter and adjacent to one another. Site layout can specify that all services be routed to the same end of the shelter. By the way, if the site has water from a commercial service or from a well, the water pipe should enter the facility at the same location and should be bonded to the single point. If the shelter has a structural lightning protection system, one of its grounding points should be bonded to the single point.

Transient suppression

If single-point grounding manages to reduce all equipment chassis and service grounds to the same electrical potential, what about the "hot" or "line" sides of those services? Any excess difference in electrical potential between the line side and the unified ground still will cause equipment damage or accelerated wear. Single-point grounding does not and cannot resolve that issue; therefore, it is important to install transient voltage surge suppression (TVSS) equipment in a "staged protection" design, with the primary TVSS device at the service entrance, to prevent equipment damage caused by an excess difference in electrical potential between the ground and hot or line sides of the services.

Lightning protection is a three-part process involving effective bonding and grounding, transient voltage surge suppression and structural lightning protection. Without all three elements, site equipment is not protected. Single-point grounding is but one element; nevertheless, it is probably the most basic element and the most critical to design into a site. The single-point grounding concept is not empty theory. It has been field-tested in the most lightning-damage-prone areas of the country and has proven to be effective. Put this simple concept to work to reduce damage and equipment wear.

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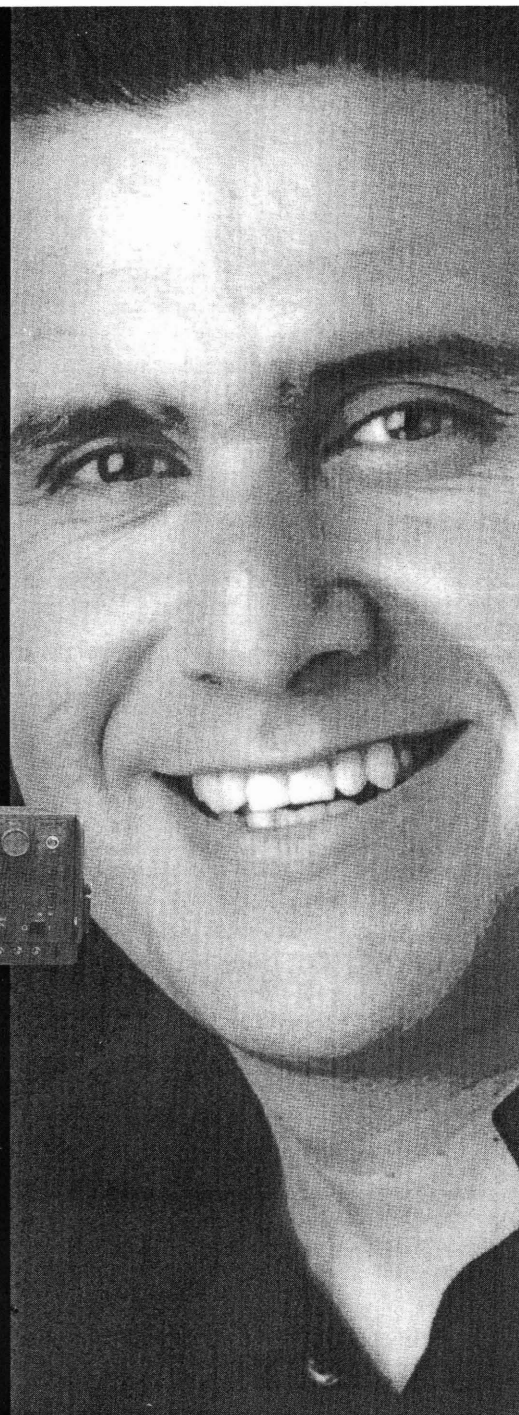
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Receiver bandwidth and the frequency-modulated signal

By Harold Kinley, CET

It is important that an FM receiver's bandwidth be wide enough to accommodate a fully deviated FM signal.

Unless the bandwidth is wide enough to accommodate the FM sidebands, severe distortion results. This distortion can cause squelch blocking on fully modulated signals.

Figure 1 below shows the instantaneous frequency excursions from the carrier frequency (f_c). Here, a 1,000Hz audio modulating tone (f_M) is

Kinley is a certified electronics technician with the South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985.

frequency-modulating the carrier at a deviation level of $\pm 5\text{kHz}$. If the receiver bandwidth were perfectly flat to $\pm 5\text{kHz}$ as shown by the box outline, it would appear that all of the modulated signal would fall within the receiver's passband. Not so.

FM spectrum

Unlike an AM signal, where, with single-tone modulation, there are only two sidebands, in an FM signal there are (theoretically) an infinite number of sidebands spaced on either side of the carrier at $f_c \pm f_M \pm 2f_M \pm 3f_M \dots$. This means that, for the FM transmission mode to be distortion-free, an infinite bandwidth is required.

Nevertheless, as a practical matter, it has been established that a low-distortion signal results with a band-

width containing only those sidebands that represent 98% of the unmodulated carrier power. Sideband pairs that represent at least 2% of the unmodulated carrier power are called the significant sidebands.

The formula known as Carson's rule for establishing the minimum required bandwidth for frequency modulation is:

$$B_W = 2(\Delta f + 1)f_M$$

where

B_W = bandwidth

Δf = maximum frequency deviation

f_M = modulating frequency

In his 1968 book *Communications Systems* published by McGraw-Hill, Bruce A. Carlson established the following formula as "... a better approximation for equipment design to determine 3dB bandwidth of RF and IF amplifiers ...":

$$B_W \approx 2(\Delta f + 2f_M)$$

Sideband power

The figures in Table 1 bottom left were taken from a Bessel function chart.

The table gives the voltage amplitude of all of the FM signal's significant sidebands relative to the unmodulated carrier-voltage amplitude when the modulation index (β) is 5.0.

The formula for finding modulation index (β) is:

$$\beta = \Delta f / f_M$$

Examine Figure 2 on page 64.

This graph illustrates the power distribution of at least 98% of the FM spectrum power given Figure 1's modulating conditions. The modulating tone (f_M) is 1,000Hz; the frequency deviation (Δf) is 5kHz; and the modulation index (β) is 5.0.

Notice in Figure 2 that significant sideband pairs extend out to $f_c \pm 6\text{kHz}$. If a receiver had a passband represented by Figure 1, the sideband

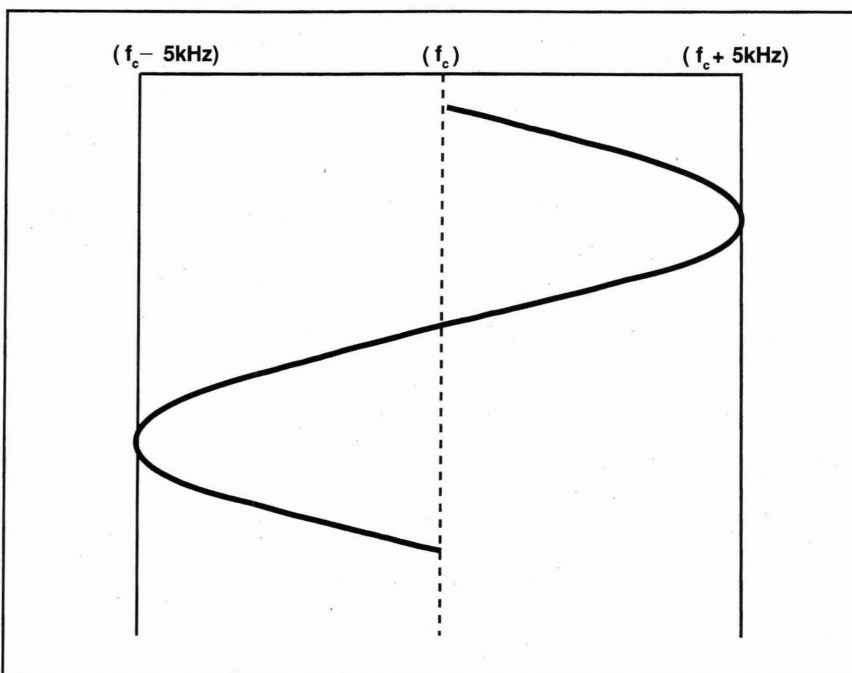


Figure 1. A 1,000Hz sine wave is frequency-modulating a carrier f_c at a deviation level of $\pm 5\text{kHz}$. The modulation index (β) is 5.0. The boxed outline represents the receiver passband. Notice that the instantaneous frequency excursions at the maximum and minimum points do not exceed the receiver's passband.

Table 1—The voltage amplitude of all of the significant sidebands of the FM signal relative to the unmodulated voltage when the modulation index is 5.0.

β	f_c	$f_c \pm f_M$	$f_c \pm 2f_M$	$f_c \pm 3f_M$	$f_c \pm 4f_M$	$f_c \pm 5f_M$	$f_c \pm 6f_M$
5	.1776	.3276	.0466	.3648	.3912	.2611	.131

(continued on page 64)

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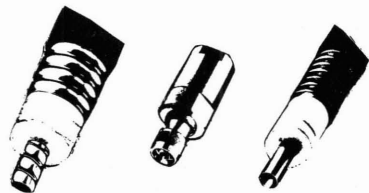
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Technically speaking

100V ACROSS 50Ω = 200W = 0dB REFERENCE

DEVIATION = ±5kHz, TONE = 1kHz, MODULATION INDEX = 5.0

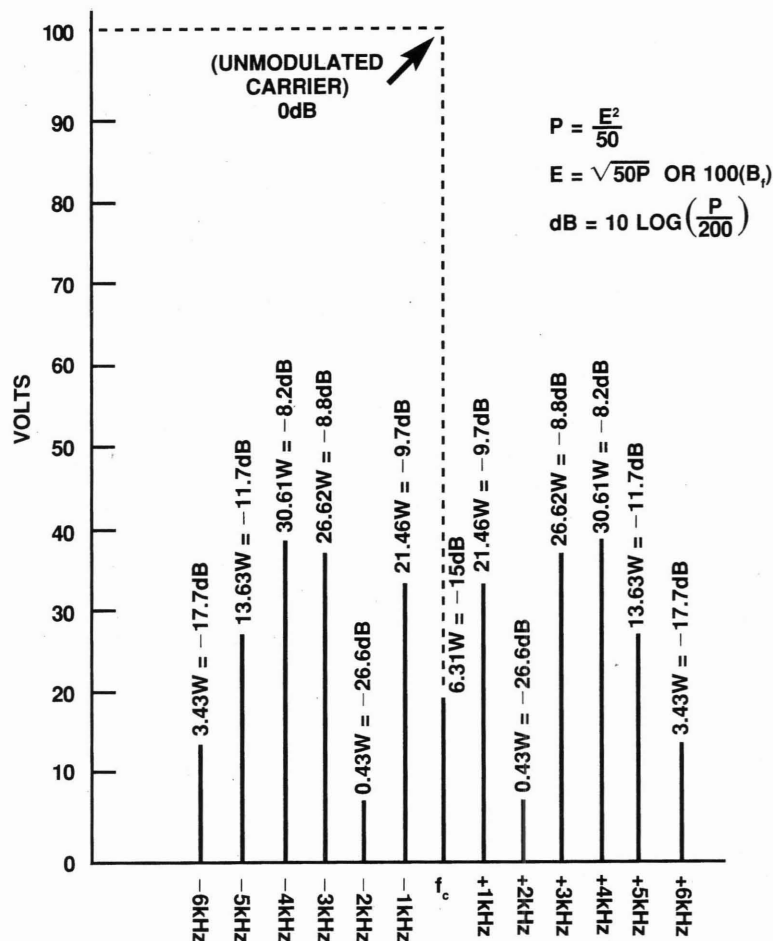


Figure 2. The frequency-modulated signal produced by the conditions depicted in Figure 1 produce the sideband spectrum shown above. Only the significant sidebands are included. Note that the carrier is down 15dB from its unmodulated level. If the deviation were $\approx \pm 5.5$ kHz with a 1kHz tone (to produce a modulation index $[\beta]$ of 5.5), the carrier would null out com-

pletely and then rise again as the modulation index changes. This phenomenon can be used to calibrate deviation meters accurately by using a spectrum analyzer to locate the carrier null points. There are many values of modulation index (β) that will produce a carrier null (2.405, 5.52, 8.65, etc.).

(continued from page 8)

pairs at ± 6 kHz would be excluded from the received signal, causing distortion.

Notice in Figure 2 that the 0dB reference level is 100V. This represents a power level of:

$$100^2 / 50 = 200W$$

The power level of the carrier and each sideband is derived by finding the appropriate Bessel function in Table 1, then substituting into this formula:

$$P_s = \frac{[E_c(B_f)]^2}{50}$$

where

Technically speaking

P_S = sideband power in watts

E_C = the unmodulated carrier voltage

B_f = Bessel function

For example, to find the second sideband's power level, $f_C + 2f_M$, find the Bessel function B_f : it is .0466. Sub-

stituting into the equation:

$$P_S = \frac{[100(.0466)]^2}{50} = 0.43W$$

All of the individual sideband power levels (through $f_C \pm 6f_M$) have been calculated as shown above and are listed in Figure 2. The decibel figures are included.

When you look at this particular signal on a spectrum analyzer display, you will notice that the individual sidebands' relative amplitudes may appear to be incorrect. Nevertheless, keep in mind that on the spectrum analyzer display, the scale is calculated in linear decibel units, typically 10dB/division.

If you first establish the 0dB unmodulated carrier reference point, you then can use the decibel figures in Figure 2 to compare with the spectrum analyzer display.

For example, you will notice that the carrier (at 5kHz deviation with a 1,000Hz tone) will be 15dB down from the unmodulated carrier (0dB reference level). When you add the power level of all of the individual sidebands plus the carrier, the total is:

$$\begin{aligned} &6.31 + 2(21.46) + 2(0.43) + 2(26.62) \\ &+ 2(30.61) + 2(13.63) + 2(3.43) \\ &= 198.67W. \end{aligned}$$

In terms of percentage of power, this figure represents:

$$P\% = 100 \left[\frac{198.67}{200} \right] = 99.33\%$$

where

$P\%$ = percentage of power in the sidebands + carrier

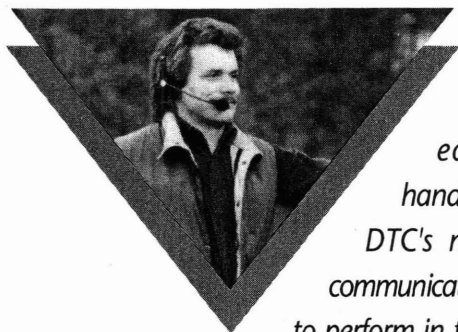
Because this is greater than 98%, all of the *significant* sidebands are included. The important point to remember is that when the carrier is frequency-modulated, the FM spectrum's total power does not change, it simply is redistributed.

If you were to connect a *very* selective 5kHz-bandwidth filter centered at f_C between the transmitter producing the spectrum shown in Figure 1 and a wattmeter, the wattmeter would read only the *sum* of the powers in the carrier and the first two sideband pairs. This would be:

$$6.31 + 2(21.46) + 2(0.43) = 50.09W$$

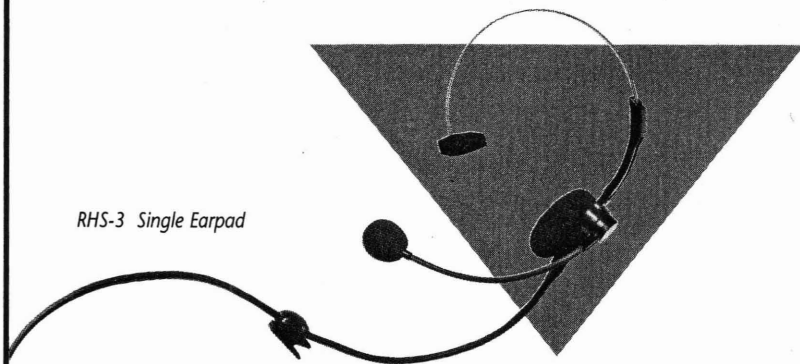
representing approximately 25% of the total spectrum power. If the modulation were removed, leaving only the carri-

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Technically speaking

er, the wattmeter then would indicate the full 200W.

Fortunately, the greatest power distribution from the speech waveform lies in the lower frequencies. For lower frequencies, the sidebands are spaced closer and thus, for a given modulation index, occupy less bandwidth.

The average modulation index of a typical complex speech waveform might be from 2 to 2.5. Compressed speech waveforms would produce a higher average modulation index than uncompressed speech waveforms.

The preceding discussion ties in directly with the measurement of the FM receiver's bandwidth. If the bandwidth is too narrow to pass the FM signal's significant sidebands, then serious signal quality degradation results. Two methods of measuring FM receiver bandpass response are described.

The term *modulation acceptance bandwidth* refers to the deviation level at which the signal is degraded by 6dB as referenced to the 12dB SINAD sensitivity input level.

For example, if the 12dB SINAD sensitivity level is $0.25\mu\text{V}$, then the input signal level is increased by 6dB to $0.5\mu\text{V}$, and then the deviation is increased from 3kHz to a level that returns the SINAD meter reading to the 12dB SINAD level. This is the 6dB bandwidth of the receiver.

This procedure for checking the modulation acceptance bandwidth is described in the Institute of Electrical and Electronics Engineers (IEEE) publication *IEEE Test Procedure for FREQUENCY-MODULATED MOBILE COMMUNICATIONS RECEIVERS* (IEEE Std. #184-1969) para. 4.1.1.1.3, p. 6. Typically, a properly operating receiver yields a test result of $\pm 7.0\text{kHz}$ to $\pm 7.5\text{kHz}$ modulation acceptance bandwidth.

Minimum usable bandwidth

The Electronic Industries Association (EIA) standard *MINIMUM STANDARDS FOR LAND MOBILE COMMUNICATION FM OR PM RECEIVERS, 25-947 MHz* (EIA-204-C) para.

8.0, p. 11, describes a slightly different method of establishing the 6dB bandwidth.

As does the IEEE method, the EIA method uses the 12dB SINAD method of establishing the 6dB bandwidth. Moreover, the EIA method uses the 12dB SINAD signal input level as a reference (with a *center-frequency* signal).

The signal generator level then is increased by 6dB. Next the signal generator frequency is *increased* until the SINAD meter again indicates 12dB SINAD. The signal generator's frequency then is recorded.

Next, the signal generator's frequency is tuned below the center frequency until the SINAD meter again indicates 12dB SINAD. The signal generator frequency is recorded.

The frequency difference (Δf) between the center frequency (f_c) and the lower frequency (f_L) and the difference between the upper frequency (f_U) and the center frequency (f_c) are calculated. The smaller difference is the *minimum usable bandwidth*.

The EIA standard further states: "The minimum usable bandwidth shall not be less than one-half of the maximum specified receiver frequency stability in kHz (but at least 40% of the rated system deviation.)"

For a system deviation of $\pm 5.0\text{kHz}$, the minimum Δf should be no less than $\pm 2.0\text{kHz}$. Remember, the generator signal is being deviated at $\pm 3.0\text{kHz}$ already, so that this deviation level has to be considered along with the frequency displacement (Δf).

The procedures described above are similar in principle but different in methodology. The modulation acceptance bandwidth test prescribed by the IEEE standard is easier to carry out on the workbench and is more commonly used by manufacturers in specifying receiver bandwidth and by technicians in running the test.

Both methods yield valid results, but the one you use should be dictated by the one used by the manufacturer in writing the specification sheet.

When receiver alignment and repairs are done, it should be standard practice to check the modulation acceptance bandwidth or minimum usable bandwidth, just as you would check the 12dB SINAD sensitivity.

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Intermod: Getting the upper hand (Part 2)

By Harold Kinley

"An ounce of prevention is worth a pound of cure" certainly applies to the problem of intermodulation interference. Several methods are used to get the upper hand on intermodulation interference. The method used to suppress the IM will depend on where the IM is generated.

Figure 1 shows an RF amplifier with two signals (*A* and *B*) ap-

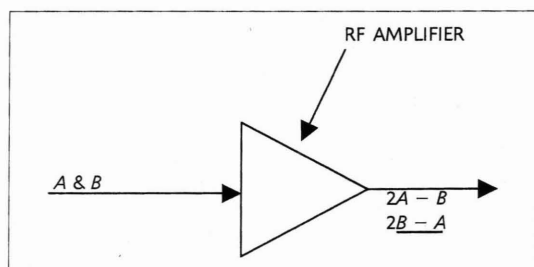


Figure 1: If two input signals, *A* and *B*, are applied to the input of an amplifier, the mixing process would produce two intermodulation products, $2A - B$ and $2B - A$. The level of the IM products at the output of the amplifier will depend on the levels of *A* and *B* at the input and the third-order intercept point of the amplifier. The TOIP is a theoretical figure and one used to quantify the intermodulation rejection ability of the amplifier. The higher this figure, the greater the intermod rejection capability of the amplifier.

Table 1: The relationship of the *A* and *B* input signal levels, the third-order intercept point and the intermod levels at the amplifier output.

Column	A	B	C	D	E	F	G
TOIP	+5	+10	+20	+20	+20	+20	+20
A	-10	-10	-10	-13	-9	+10	+20
B	-10	-10	-10	-13	-9	+10	+20
$2A - B$	-40	-50	-70	-79	-67	-10	+20
$2B - A$	-40	-50	-70	-79	-67	-10	+20

plied to the input. Third-order intermod signals, $2A - B$ and $2B - A$, will appear at the output of the amplifier. Other products will appear at the output as well. However, this discussion will be limited to the third-order IM products. The level of the third-order IM products at the ampli-

fier output will depend on the level of the *A* and *B* signals at the input and the third-order intercept point (TOIP) of the amplifier.

Table 1 illustrates the relationship between input levels: the TOIP and IM levels at the amplifier output. In column A, the amplifier's TOIP is +5dBm. This is a low-quality amplifier. Column A further indicates that if the input levels are at -10dBm, the IM levels at the amplifier output are at -40dBm. In column B, the input levels are the same as column A; but, the TOIP of the amplifier is +10dBm and the IM levels at the amplifier output have dropped to -50dBm. Note that the drop in the IM level is equal to twice the amount of the increase in the TOIP. In column C, the TOIP increases by 10dB over column B and the IM level in column C drops by 20dB.

This illustrates the importance of using an amplifier with a high TOIP. Amplifiers with TOIPs much higher than this are used in situations where greater IM suppression is a must. However, these higher-quality amplifiers are accompanied by a higher price tag. It's still true—you don't get something for nothing. Install a cheap amplifier in front of a receiver in a densely populated site, and

you will hear things you have never heard before—and wish you wouldn't hear.

If you are lucky enough to have a site where an amplifier can help, then use the highest quality available. Though you might get away with using a lower-quality amplifier today, you probably

won't for long.

Refer again to Table 1. Column D shows the two input signals at -13dBm. That is 3dB down from column C. Yet, the output levels of the IM signals are 9dB down from column C. Thus, we have realized a three-fold reduction in the IM signal level compared to the input signals. This is a characteristic of the IM mixing process. For third-order IM, a reduction of 1dB in each of the input signals (*A* and *B*) causes a drop of 3dB in the IM signals. So, for third-order IM we get a three-for-one advantage. That is, for every decibel of attenuation in the input signals, we get a 3dB reduction in the IM product. With fifth-order products, we get a five-for-one reduction. This can be generalized as such: A reduction of 1dB to the intermod-forming input signals will yield *N* decibels of reduction to IM signal for an *N*th-order IM signal. This reduction must occur ahead of the mixing point. This is an important characteristic—one that can be used to our advantage in suppressing IM levels.

In Table 1, examine column E. Here, the input signal levels are increased by 1dB compared to column C (from -10 to -9dBm). Yet, the IM signals at the output have increased by 3dB. This is the reverse of the above situation where the input signals were reduced. This still complies with the IM mixing rules. A change in the levels of the IM forming signals ahead of the mixing point will result in a greater change in the IM signal by an amount equal to the order of the

Contributing editor Kinley, MRT's technical consultant and a certified electronics technician, is regional communications manager, South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual, with Instrumentation and Testing Techniques*, which is available for direct purchase. Write to 204 Tanglewylde Drive, Spartanburg, SC 29301. His email address is halkinley@charter.net.

IM signal.

Taking a final look at Table 1, examine column G. Here, the input signals are equal to the TOIP of the amplifier and the IM signals are equal to the input signal levels. This theoretical point can never be reached in practice be-

cause the amplifier would become practically inoperative before this point is reached.

Transmitter-produced IM

The class C output stage of a transmitter is fertile ground for the production of intermod. By design,

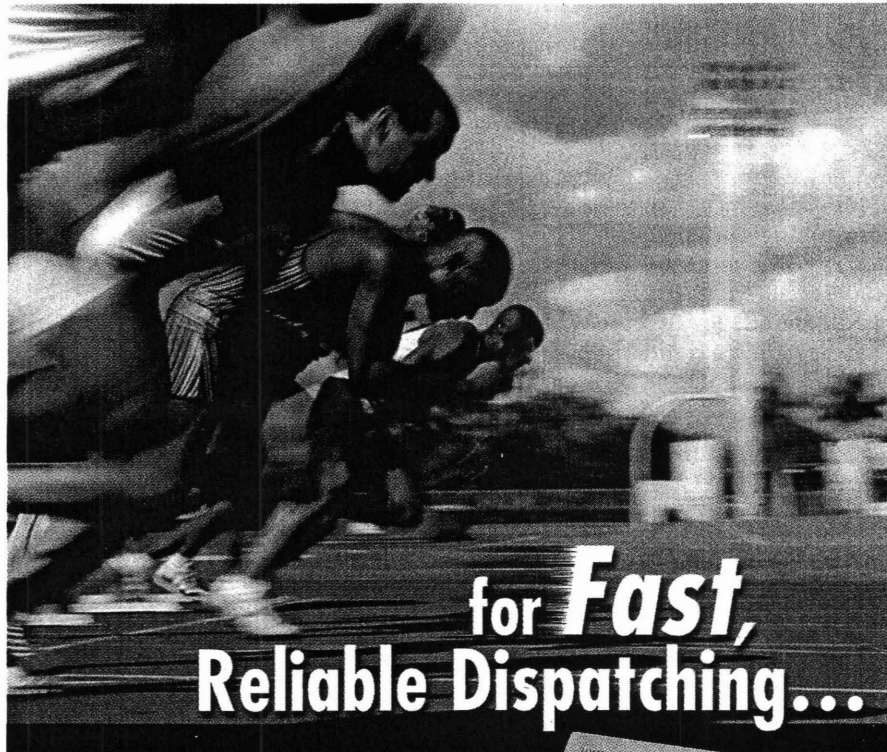
it is non-linear, rich in harmonics and connected to an antenna. Let's look at a typical two-signal IM product of the third order that is capable of interfering with a nearby receiver. We will assign some frequencies and calculate an approximate level for the IM signal. Then we will apply some solutions and check the final outcome.

Figure 2 shows two transmitters: A and B. Receiver C is operating at a frequency that is equal to $2A - B$. The signal from transmitter B enters the final stage of transmitter A where it mixes with the signal A to form the $2A - B$ IM signal. This is on the frequency of receiver C and therefore causes interference to the receiver. The calculation of the level of the IM signal is shown at the right. Because the level of the IM signal is -79dBm , it will seriously degrade the performance of the receiver. Steps must be taken to suppress the IM signal to a non-interfering level. The amount of suppression necessary will depend on the site noise level and the minimum necessary receive level.

Figure 3 shows that to suppress the IM signal to a non-interfering level, a bandpass cavity filter and an isolator have been installed on transmitter A where the IM signal is produced. Figure 4 shows the selectivity curve of the bandpass filter. Note that as the signal from transmitter B passes through the filter, it is attenuated by 20dB. The isolator offers another 35dB of attenuation to signal B before it reaches the final amplifier stage of transmitter A.

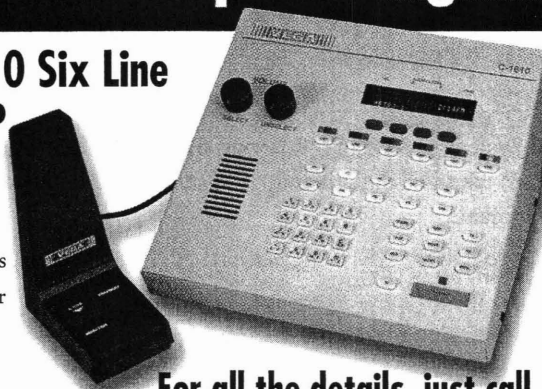
Still, an IM signal is formed in transmitter A and travels back up the line through the isolator with negligible attenuation. But in passing back through the bandpass cavity filter, the IM signal is attenuated by 20dB. This means that the IM signal leaving the antenna and reaching receiver C is 75dB down from what it was before the isolator and cavity filter were installed.

In this case, we did not gain any leverage in reducing the IM signal. To realize any leverage, both signals, or at least the signal with the



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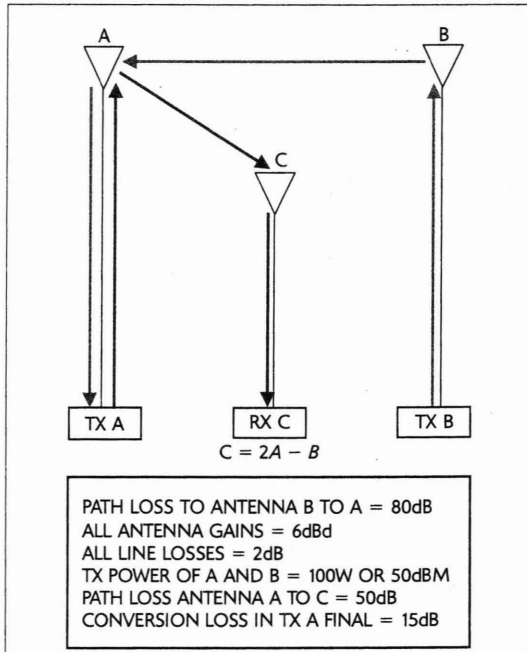


Figure 2: The signal from TX B (shown in green) enters the final stage of TX A, where it mixes with the signal from TX A to form the intermod signal, $2A - B$ (shown in red). This IM signal, at the frequency of RX C, causes interference to RX C. The IM level at the input of RX C is -79dBm . See below for calculations.

TRANSMITTER B	+50dBm
LINE LOSS B	-02
ANT. GAIN B	+06
PATH LOSS ANT. B TO A	-80
ANT. GAIN A	+06
LINE LOSS A	-02
CONVERSION LOSS TX A	-15
LINE LOSS A	-02
ANT. GAIN A	+06
PATH LOSS ANT A TO C	-50
ANT. GAIN C	+06
LINE LOSS C	-02
INPUT TO RX C	-79dBm

coefficient greater than one, must pass through some attenuation ahead of the mixing point. Another mixing point that could generate the $2A - B$ IM signal is in transmitter B. However, the IM generated there would be at a much lower level because of the leverage effect. This is because the A signal would be attenuated by the path loss between the two transmitters. Because this attenuation would be leveraged by a factor of two, the resulting IM signal in transmitter B would be of such a

low level as to be inconsequential.

Receiver-produced intermod

There is a third point in Figure 2 where the $2A - B$ IM signal could form. It is in the receiver itself. If the receiver front end were sufficiently overloaded, it would become nonlinear and become a good mixing point for the production of IM products. (See Figure 5.) In this situation, it is possible to get leverage from suppressing the IM signals. Remember, if both IM forming signals (A and B) are attenuated prior to the mixing point, the amount of reduction in the IM level is leveraged by a factor equal to the sum of the coefficients of the individual signals, A and B.

In Figure 6, an attenuator is placed between the antenna and the receiver input so that the individual signals, A and B, must pass through the attenuator. If the attenuator is set to 3dB, then we will realize a 9dB reduction in the $2A - B$ IM signal. This gives us real leverage in dealing with the IM problem. Because the desired signal must also pass through the attenuator, it will also be attenuated by 3dB. So the net gain, in terms of carrier-to-interference ratio, will be 6dB. In a situation where the IM is of low amplitude and the site noise is high, a simple resistive attenuator might be all that is needed to resolve the problem.

Usually, the cure is not that simple. It becomes a matter of selectivity ahead of the receiver. The more selectivity that exists ahead

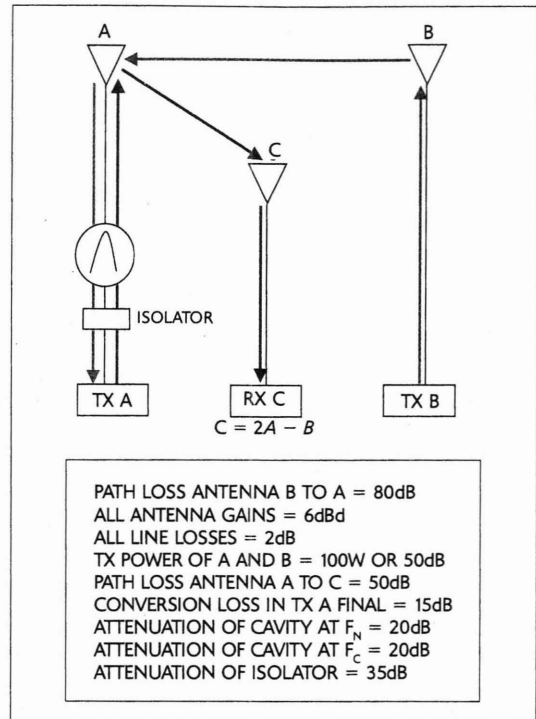


Figure 3: An isolator and a bandpass cavity have been added in the line at TX A. This provides an additional 75dB of suppression of the intermod signal. This places the intermod signal at the input of RX C at a level of -154dBm . This is well below the receiver sensitivity threshold. Thus, the IM is dead and buried. See below for the calculations.

TRANSMITTER B	+50dBm
LINE LOSS B	-02
ANT. GAIN B	+06
PATH LOSS ANT. B TO A	-80
ANT. GAIN A	+06
LINE LOSS A	-02
CONVERSION LOSS TX A	-15
LINE LOSS A	-02
ANT. GAIN A	+06
PATH LOSS ANT A TO C	-50
ANT. GAIN C	+06
LINE LOSS C	-02
CAVITY ATTENUATION TX B	-20
ISOLATOR ATTENUATION	-35
CAVITY ATTENUATION TO IM	-20
INPUT TO RX C	-154dBm

of the first active receiver stage, the better is the IM rejection capability of the receiver. It may be necessary to connect two or three bandpass cavity filters in cascade to achieve sufficient suppression of the IM signal. (See Figure 7.) This bandpass response curve represents a bandpass filter arrangement placed in front of the

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Technically Speaking

receiver. The response of the filter is such that the attenuation at the desired receiver frequency (154MHz) is 2dB. The attenuation at frequency A (155MHz) is 10dB. This will be leveraged by a factor of two from the IM form of $2A - B$. The attenuation at frequency B (156MHz) is 25dB. Thus the total reduction in the IM signal in the receiver is:

$$2(10) + 25 = 45\text{dB}$$

Another possible filter method uses a notch filter. If it is desired to notch out both signals, A and B, two notch filters would be required. If we can only use a single notch filter, we have to choose which of the two IM-forming signals to notch. It would be best to notch out the A signal (155MHz) because a leverage factor of two would be realized. If we choose to notch out the B signal (156MHz) no leverage would be realized and we would only achieve a decibel for decibel reduction in the IM signal. To notch out both of these signals would require two notch cavities.

The advantage of using notch cavities is that more attenuation (at a specific frequency close to the desired frequency) can be achieved than with bandpass cavities. The disadvantage is that the notch filter only helps at one frequency while the bandpass filter helps at many frequencies outside the passband of the filter. The choice of which type of filter to use will depend on the degree of suppression needed and the spacing relative to the desired receive signal.

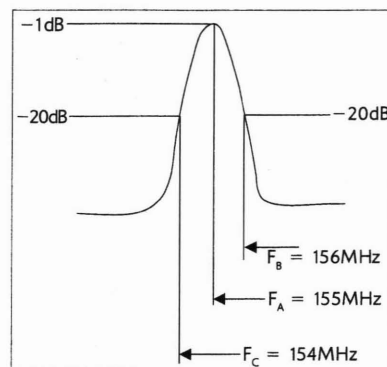


Figure 4: A cavity filter with this selectivity characteristic is placed in the TX A line. The center frequency at 155MHz passes the normal TX A signal with little attenuation. The attenuation at the frequency of TX B is 20dB, and the attenuation at RX C frequency (the intermod signal) is 20dB. So, this filter provides dual action to suppress the IM signal.

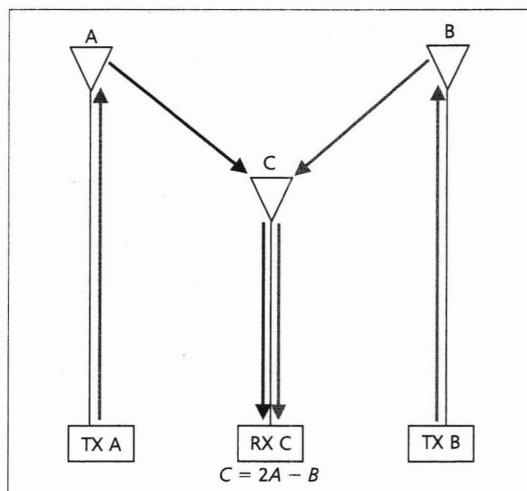


Figure 5: The signal from TX A (shown in blue) and the signal from TX B (shown in green) enter the front end of RX C where they mix to form the intermod signal, $2A - B$. This IM signal, at the frequency of RX C, causes severe interference to RX C.

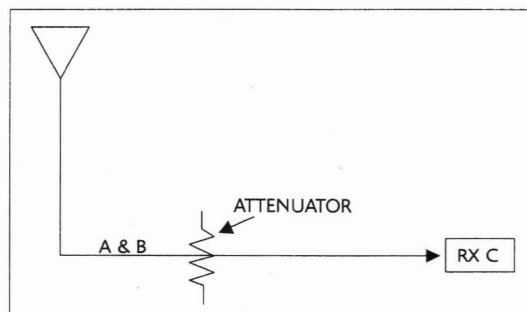


Figure 6: A simple attenuator is placed ahead of the receiver so that the A and B signals must pass through it before getting into the receiver. This simple attenuator can provide leverage in suppressing the IM signal.

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Technically Speaking

Crystal filters are now available for operation at VHF highband range and can provide excellent selectivity. Typically, the crystal filter has a comparatively high insertion loss at the desired pass frequency. Some IM generation is also inherent in crystal filters, but the good usually outweighs the bad, yielding a positive net result.

Many times interference is called intermod when it really isn't intermod. Spurious signals are not always the result of intermodulation. One telltale sign of intermod is a signal that might be heard in the middle of a conversation then cut off abruptly. Another sign is an over-deviated signal. The key to suppressing the IM signal is identifying the mixing point and the individual component signals that form the IM. It is possible to have a combination of mixing points. For example, there might be a combination of transmitter-produced IM and receiver-produced IM. When suppression techniques are applied to the transmitter, the results might first appear promising. Then, further efforts might not yield further results. If the IM from the transmitter is suppressed below the level of the IM produced in the receiver, the receiver-produced IM will become the dominant IM and suppression techniques must be applied at the receiver.

For further reading, I highly recommend *Intermod Control* by William F. Lieske Sr., founder and retired president of EMR. I con-

sider it the definitive work on intermodulation interference.

Until next time—stay tuned! ■

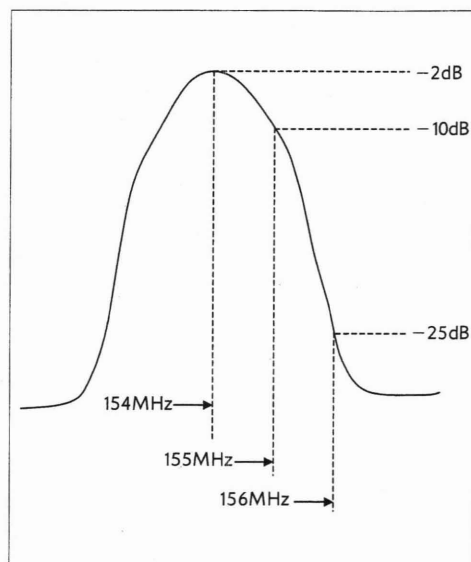


Figure 7: Two or three bandpass cavities connected in cascade could provide a bandpass selectivity such as this to suppress the IM signal level in the receiver. Note that the normal receiver frequency is placed in the center of the passband of the filter to offer little attenuation to the desired signal. The A signal is attenuated by 10dB and this will be leveraged by a factor of two while the B signal is attenuated by 25dB. This results in a 45dB reduction in the IM signal produced in the receiver.

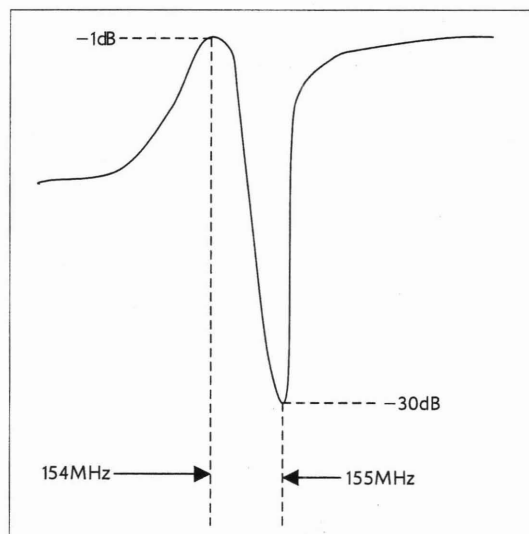


Figure 8: A notch filter such as this can be used to suppress the IM signal. The desired receiver signal at 154MHz is attenuated only 1dB, while the A signal at 155MHz is attenuated by 30dB. This 30dB is leveraged by a factor of two so the result is a 60dB reduction in the IM signal.

Demystifying error-control coding

It's the key to designing error-free communications systems. But what is it?

Part 1 of 4

Many of us have heard the term error-control coding in connection with modern land mobile radio systems, but few of us understand what it is, what it does and why we should be concerned about it. Old timers are forewarned: All LMRs eventually will be digital, and all digital radios use error-control coding. Terms like Golay code, Hamming code, CRC and interleaving will become popular topics at cocktail parties. Don't get left out of the conversation; learn about error-control codes.

This article begins a series that will examine what error-control codes do, how they work, and when they can be used successfully.

Error-control coding is a discipline under the branch of applied mathematics called Information Theory, discovered by Claude Shannon in 1948 [1]. Prior to this discovery, conventional wisdom said that channel noise prevented error-free communications. Shannon proved otherwise when he showed that channel noise limits the transmission rate, not the error probability.

Shannon showed that every communications channel has a capacity, C (measured in bits per second), and as

long as the transmission rate, R (also in bits per second), is less than C , it is possible to design a virtually error-free communications system using error-control codes. Shannon's contribution was to prove the existence of such codes. He did not tell us how to find them.

After the publication of Shannon's famous paper, researchers scrambled to find codes that would produce the very small probability of error that he predicted. Progress was disappointing in the 1950s when only a few weak codes were found.

In the 1960s, the field split between the algebraists, who concentrated on a class of codes called block codes, and the probabilists, who were concerned with understanding encoding and decoding as a random process. The probabilists eventually discovered a second class of codes, called convolutional codes, and designed powerful decoders for them.

In the 1970s, the two research paths merged, and several efficient decoding algorithms were developed. With the advent of inexpensive microelectronics, decoders finally became practical and in 1981, the entertainment industry adopted a very powerful error control scheme for the new CD player [2]. Today, error-control coding in its many forms is used in almost every new communications system, including the As-

sociation of Public-Safety Communications Officials' Project 25 standard.

Digital communications systems often are conceptualized as shown in Figure 1. The following paragraphs describe the elements of Figure 1 and define other terms common to error-control coding.

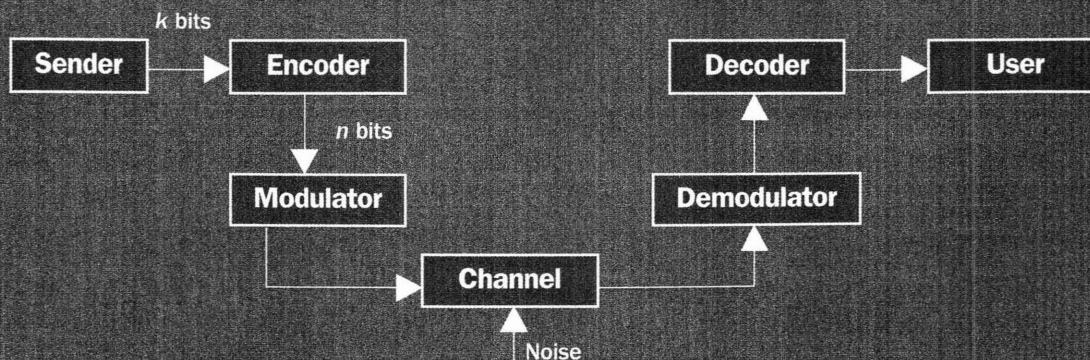
■ **Encoder and decoder:** The encoder adds redundant bits to the sender's bit stream to create a code word. The decoder uses the redundant bits to detect and/or correct as many bit errors as the particular error-control code will allow. For our purposes, encoding and decoding refers to channel coding as opposed to source coding. (Source coding prepares the original information for transmission; the vocoder, or voice-encoder, is one example of source coding.)

■ **Modulator and demodulator:** The modulator transforms the output of the encoder, which is digital, into a format suitable for the channel, which is usually analog (e.g., a radio channel). The demodulator attempts to recover the correct channel symbol in the presence of noise.

When the wrong symbol is selected, the decoder tries to correct any errors that result. Some demodulators make soft decisions, meaning that the de-

Continued on page 34

Figure 1



Continued from page 30

modulator does not attempt to match the received signal to one of the allowed symbols. Instead, it matches the noisy sample to a larger set of discrete symbols and sends it to the decoder where the heavy lifting is done.

All LMRs eventually will be digital, and all digital radios use error-control coding.

■ **Communications channel:** This part of the communication system introduces errors. The channel can be radio, twisted wire pair, coaxial cable, fiber-optic cable, magnetic tape, optical discs or any other noisy medium.

■ **Error-control code:** The set of code words used with an encoder and decoder to detect errors, correct errors, or both detect and correct errors.

■ **Bit-error rate (BER):** The probability of bit error often is the figure of merit for an error-control code. We want to keep this number small, typically less than 10^{-4} for data and less than 10^{-3} for digital voice. BER is a useful indicator of system performance on an independent error channel, but it has little meaning on bursty, or dependent, error channels.

■ **Message-error rate:** This probability of message error is sometimes called frame error rate. This may be a more appropriate figure of merit because the smart user wants all of his or her messages to be error-free and couldn't care less about the BER.

■ **Undetected message error rate (UMER):** This is the probability that the error detection decoder fails and an errored message (code word) slips through undetected. This event happens when the error pattern introduced by the channel is such that the transmitted code word is converted into another valid code word. The decoder can't tell the difference and must conclude that the message is error-free. Practical error

detection codes ensure that the UMER is very small, often less than 10^{-16} .

■ **Random errors:** These errors occur independently. This type of error occurs on channels that are impaired solely by thermal (Gaussian) noise. Independent-error channels also are

called memory-less channels because knowledge of previous channel symbols adds nothing to our knowledge of the current channel symbol.

■ **Burst errors:** These errors are not independent. For example, channels with deep fades experience errors that occur in bursts. Because the fades result in consecutive bits that are more likely to

Equation 1

$$E_b = \frac{P_t}{R}$$

be in error, the errors are usually considered dependent rather than independent. In contrast to independent-error channels, burst-error channels have memory.

■ **Energy per bit:** This refers to the amount of energy contained in one information bit. This is not a parameter that can be measured by a meter, but it can be derived from other known parameters. Energy per bit (E_b) is important because almost all channel impairments can be overcome by increasing it. Energy per bit (in joules) is related to transmitter power P_t (in watts and bit rate R (in bits per second), as shown in Equation 1.

If transmit power is fixed, the energy

per bit can be increased by lowering the bit rate. This is why lower bit rates are considered more robust. The required energy per bit to maintain reliable communications can be decreased through error-control coding, as we shall see in the next article in this series.

■ **Coding gain:** This refers to the difference in decibels (dB) in the signal-to-noise ratio required to maintain reliable communications after coding is employed. Signal-to-noise ratio is usually represented as E_b/N_0 , where N_0 is the noise power spectral density measured in watts/Hertz (joules). For example, let's say a communications system requires an E_b/N_0 of 12 dB to maintain a BER of 10^{-5} , but after coding it requires only 9 dB to maintain the same BER. In that case, the coding gain is $12 \text{ dB} - 9 \text{ dB} = 3 \text{ dB}$. (Recall that $\text{dB} = 10 \log_{10} X$, where X is a ratio of powers or energies.)

■ **Code rate:** Consider an encoder that takes k information bits and adds r redundant bits (also called parity bits) for a total of $n = k + r$ bits per code word. The code rate is the fraction k/n , and the code is called an (n, k) error-control code. The added parity bits are a burden (i.e., overhead) to the communications system, so the system designer often chooses a code for its ability to achieve high coding gain with few parity bits.

Next month: What coding can and cannot do. ■

Jay Jacobsmeier is president of Pericle Communications Co., a consulting engineering firm located in Colorado Springs, Colo. He holds bachelor's and master's degrees in Electrical Engineering from Virginia Tech and Cornell University, respectively, and has more than 20 years experience as a radio frequency engineer.

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- [1] C. E. Shannon, "A Mathematical Theory of Communication," Bell System Technical Journal, vol. 27, pp. 379-423, 1948.
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Error-control coding for P25 radios

Adapt coding schemes to channel characteristics to maximize performance

Part 4 of 4

Project 25 is a digital radio airlink standard designed to create interoperability on 12.5 kHz and 6.25 kHz channels. Although Project 25 radios can pass voice or data communications, this article will focus strictly on error-control coding for voice communications.

P25 modulation comes in two flavors, Compatible 4 FM (C4FM) and Compatible Quadrature Phase Shift Keying (CQPSK) [1]. C4FM is intended for Phase I P25 systems that operate on 12.5 kHz channels, while CQPSK is intended for Phase II P25 systems that will operate on 6.25 kHz channels. In both cases, the signaling rate is 4.8 k symbols per second (sps) and there are two bits per symbol for a gross bit rate of 9.6 kb/s.

The vocoder (voice encoder/decoder) adopted by the APCO Project 25 committee is the Improved Multi-Band Excitation (IMBE) vocoder, developed by Digital Voice Systems Inc. [2]. The IMBE vocoder operates at a basic rate of 4.4 kb/s, but an additional 2.8 kb/s is required for error-control coding. Thus, the bit rate for the vocoder and its error-control coding is 7.2 kb/s. An additional 2.4 kb/s is used for channel signaling, so the gross bit rate is 9.6 kb/s.

Rather than think in terms of bit rate, we also can think in terms of bits per vocoder frame. Each frame is 20 milliseconds (ms) long (50 frames per second) and contains 88 vocoder bits, 56 error-control-coding bits and 48 signaling bits for a total of 192 bits.

A full explanation of the IMBE vocoder is beyond the scope of this article, but suffice it to say that unlike most cell phone vocoders, IMBE was optimized for low bit-rate applications (less than

4.8 kb/s). However, like most cell phone vocoders, IMBE is model-based, meaning that rather than sending digital samples of voice, the vocoder sends the parameters to be used in a human speech model.

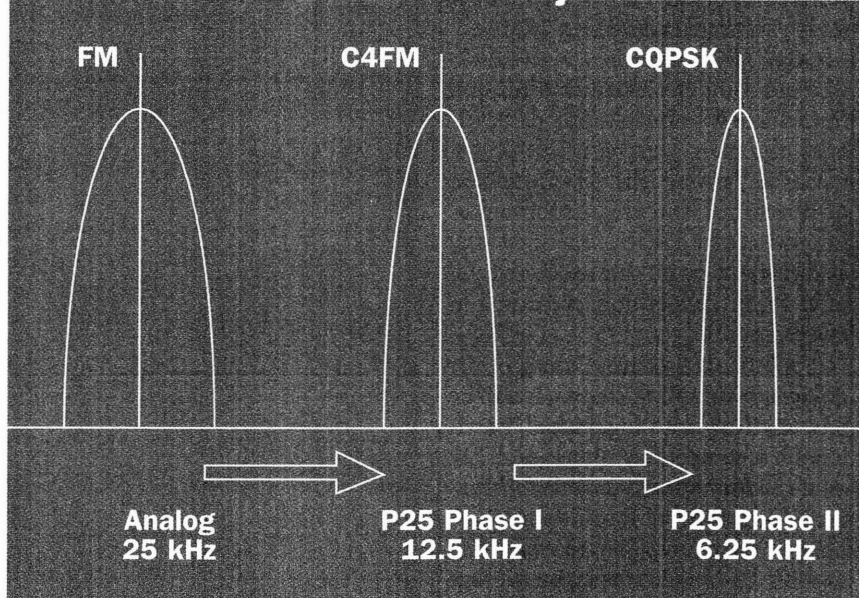
Fewer bits are needed to specify the model parameters, and both the transmitter and the receiver include a full description of the model before they leave the factory. The IMBE vocoder and its successor, AMBE, also are used

and interleaving (which can be used with any code).

The protection provided by the 56 error-control-coding bits is not applied equally over the 88 vocoder bits. Some vocoder bits are more important for intelligibility than others, so nearly all modern vocoders, including IMBE, apply error-control coding unequally. IMBE uses three principal coding techniques: interleaving, Hamming codes, and Golay codes. The signaling proto-

Figure 1

Evolution of Project 25

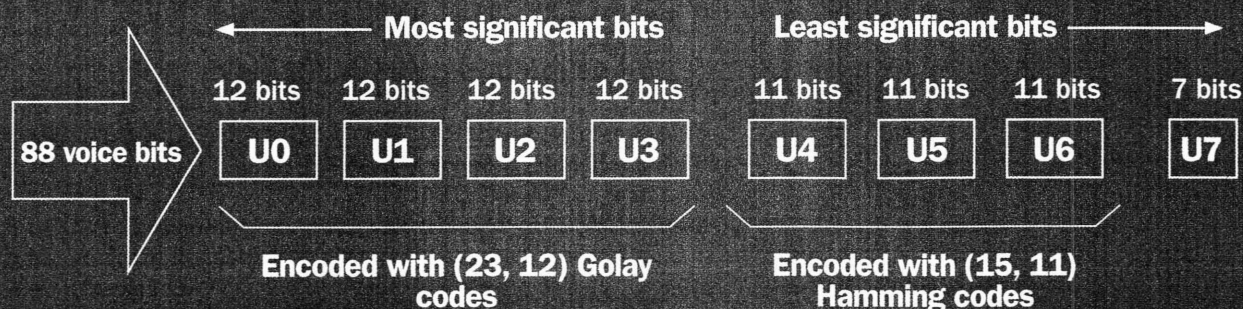


in many satellite voice networks, including Iridium's.

The mobile radio channel is hostile to digital communications, and for best performance, the error-control-coding scheme must be adapted to the channel's characteristics. Because of multipath fading, the mobile radio channel is inherently a burst-error channel, and compatible correcting techniques should be employed. Candidate burst-error-correcting techniques include Reed-Solomon codes

col (56 bits per frame) employs other coding techniques, including Reed-Solomon codes and trellis coded modulation. We'll limit the following discussion to the IMBE techniques: interleaving, Hamming and Golay codes.

Bit interleaving is used to spread short bursts of errors among several code words. The error-correction decoder is less likely to be overwhelmed if these bursts of errors arrive in multiple code blocks. The

Figure 2**Bit arrangement in 20 ms vocoder frame**

After encoding: 144 bits/frame x 50 frame/s = 7.2 kb/s

frame is interleaved in such a way that the minimum separation between any two bits of the same error-correction code is 3 symbols, or 6 bits [2]. Long bursts, such as those that occur when the user is walking or driving slowly, will be longer than the interleaver depth, so the interleaver design is not perfect. The designer is forced to make tradeoffs because greater interleaver depth means longer delay, and very long delays cannot be tolerated, especially on voice networks.

The 56 error-control bits are divided between four (23, 12) Golay codes and three (15, 11) Hamming codes. Both codes are binary block codes. Recall that an (n, k) binary block code has k information bits and $n-k$ parity bits. In our case, there are 4×11 parity bits for the Golay codes plus 3×4 parity bits for the Hamming codes, for a total of 56 error-control bits. Figure 2 shows the arrangement of bits in the vocoder frame.

Hamming codes can correct just one error per code word. They often are used in a shortened form in other communication systems for error detection and are called cyclic redundancy checks, but in P25, the Hamming code is used strictly for error correction. The Hamming code is a single error-correcting perfect code.

The (23, 12) Golay code has some unique properties. It is the only known multiple-error-correcting binary perfect code. It can correct any combina-

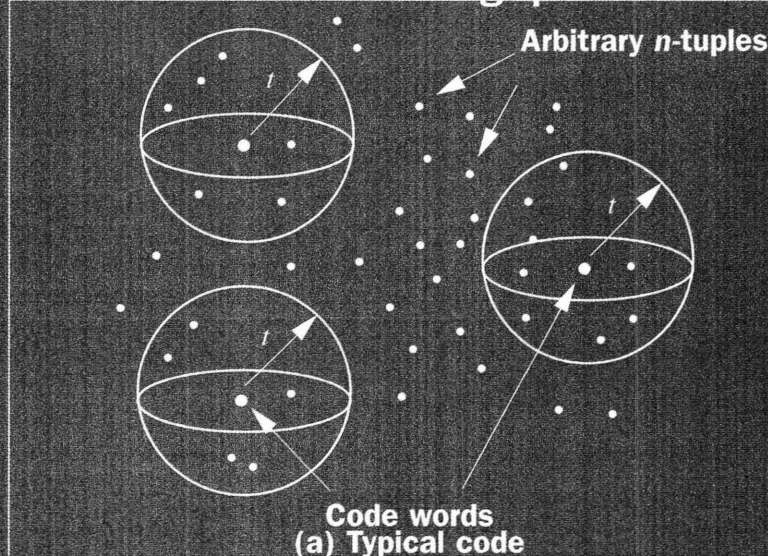
tion of three or fewer random errors in a block of 23 bits.

So, what is a perfect code? Recall from January's Tech Speak that we can represent a code word in decoding space as the center of a sphere of radius t , where t is the number of errors the code can correct. The encoder takes k bits at a time and encodes them into an n -bit code word. For a binary code, there are 2^k code words, but there are 2^n possible n -tuples to consider at the decoder because bit errors may occur on the channel.

For nearly all codes, many of the n -tuples will fall outside any sphere, and

this information can be used to detect an errored code word even if it cannot be corrected. For the very small class of perfect codes (there are only two non-trivial binary perfect codes), all n -tuples fall within a decoding sphere. This relationship is shown in Figure 3.

If this were a data system, the use of error-control codes would be relatively straightforward. Some number of parity bits would be dedicated to error detection to ensure a small probability of undetected errors. The remaining parity bits would be used for error correction. Those frames with uncorrectable error patterns (as determined by the error-detection code) would be re-

Figure 3**Decoding spheres for most**

Equation 1

$$\varepsilon_T = \sum_{i=0}^6 \varepsilon_i$$

number of errors in each received frame by computing the number of errors corrected by each of the (23, 12) Golay codes and each of the (15, 11) Hamming codes (a total of 7 code words per frame). The expression for the error estimate is found in reference [2] on page 45 or in Equations 1 and 2. It includes an estimate of the number of errors in the current frame plus a weighted version of the error estimate of the previous

different decoding sphere. This is an example of decoding failure, and it affects the accuracy of the error estimate. When the error rate is high, the probability of decoding failure also is high, and the error estimate will be somewhat flawed.

Once the error estimate is computed, the decoder does the following:

1. If the error estimate is below some threshold, it decodes the frame normally.

2. If the error estimate is above the threshold, it:

- repeats the frame,
- mutes the frame or
- performs adaptive smoothing of the frame.

When a frame is repeated, the assumption is that it is better to "stretch" the speech fragment rather than deliver a frame that is known to have problems. Remember that a frame is only 20 ms long, so a single repeat may not be noticeable to the human ear.

In a severe bit-error environment, a frame is muted. When muted, a replica of white noise is created, sometimes called "comfort noise."

Adaptive smoothing allows the decoder to take information from the previous frame and the next frame (there is a two-frame look-ahead in IMBE) to "smooth" the effects of the errored frame. Adaptive smoothing only is done over an intermediate range of bit error estimates. ■

Equation 2

$$\varepsilon_R(0) = .95\varepsilon_R(-1) + .000365\varepsilon_T$$

where $\varepsilon_R(0)$ is the estimate of the error rate for the current frame and $\varepsilon_R(-1)$ is the estimate for the previous frame.

transmitted using some form of automatic repeat request (ARQ), an error-correcting technique that always delivers error-free code words, but at the cost of decreased throughput.

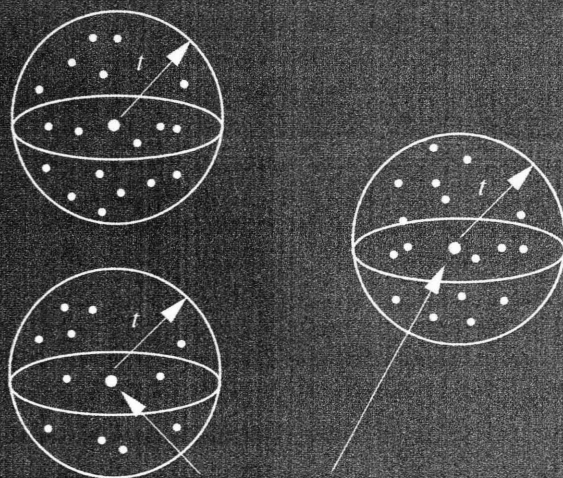
Voice systems are different. Because voice quality is subjective, we have another bag of tricks to mitigate the effect of bit errors.

The IMBE vocoder performs error estimation as part of the decoding process. The speech decoder estimates the

frame. Thus, the algorithm looks back at earlier frames and has some ability to detect trends.

When we say the number of errors corrected, we really mean the number of errors the decoder thinks it corrected. For example, if the received Golay n -tuple differs from the nearest valid code word in 1 bit position, the decoder must assume there is just one error. However, the code word may actually have five errors, placing it in an entirely

codes and perfect codes



Code words
(b) Perfect code

Jay Jacobsmeier is president of Pericle Communications Co., a consulting engineering firm located in Colorado Springs, Colo. He holds bachelor's and master's degrees in Electrical Engineering from Virginia Tech and Cornell University, respectively, and has more than 20 years experience as a radio frequency engineer.

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Error-control coding options abound

Advanced techniques for modern land mobile radios offer plusses, minuses

Part 3 of 4

Last month we covered applications and limitations of error-control codes and some of the theory behind their operation. This month we cover coding techniques. Specifically, we will discuss automatic repeat request, or ARQ, forward error correction, hybrid ARQ, interleaving and concatenation.

An error-detection code by itself does not control errors, but it can be used to request repeated transmission of errored code words until they are received error-free. This technique is called ARQ. In terms of error performance, ARQ outperforms forward error correction (FEC) because code words always are delivered error-free (provided the error-detection code doesn't fail). However, this performance does not come free of charge—we pay for it with decreased throughput.

The chief advantage of ARQ is that error detection requires simpler decoding than error correction. ARQ also is adaptive because it only re-transmits information when errors occur. On the other hand, ARQ schemes require a feedback path that may not be available.

There are two types of ARQ:

■ **Stop-and-wait ARQ:** With stop-and-wait ARQ, the transmitter sends a single code word and waits for a positive acknowledgement (ACK) or negative acknowledgement (NAK) before sending any more code words. The advantage of stop-and-wait ARQ is that it only requires a half-duplex channel. The main disadvantage is that it wastes time waiting for ACKs, resulting in low throughput.

■ **Continuous ARQ:** Continuous ARQ requires a full duplex channel because code words are sent continuously until a NAK is received. Typically, only the code word corresponding to the NAK is re-transmitted. Continuous

ARQ offers greater throughput efficiency than stop-and-wait ARQ at the cost of greater memory requirements.

FEC is appropriate for applications where the user must get the message right the first time. A voice circuit is one example. Today's error-correction codes fall into two categories:

■ **Block codes:** The repetition code introduced in last month's column (MRT, January, page 51) is an example of a binary block code. It is important to note that not all block codes are binary. In fact, one of the most popular block codes is the Reed-Solomon code that operates on m -bit symbols, not bits. Because Reed-Solomon codes correct symbol errors rather than bit errors, they are very effective at correcting burst errors. For example, a 2-symbol error-correcting Reed-Solomon code with 8-bit symbols can correct all bursts of

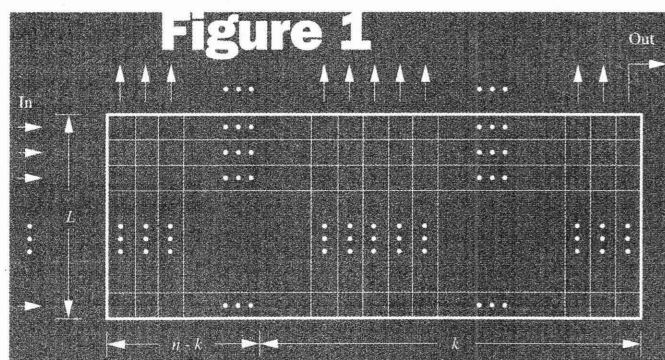
16 bits or less in length. Reed-Solomon codes are used in JTIDS, a NASA deep-space standard, and CD players.

■ **Convolutional codes:** With convolutional codes, the incoming bit stream is applied to a K -bit-long shift register. For each modification to the shift register, b new bits are inserted and n code bits are delivered, so the code rate is b/n . The power of a convolutional code is a function of its constraint length, K . Large constraint-length codes tend to be more powerful.

Unfortunately, with large constraint length comes greater decoder complexity. There are several effective decoding algorithms for convolutional codes, but the most popular is the Viterbi

algorithm, discovered by Andrew Viterbi in 1967.

One drawback of the codes we have examined so far is that they require bandwidth expansion to accommodate the added parity bits should the user wish to maintain the original unencoded information rate. In 1976, Gottfried Ungerboeck discovered a class of codes that integrates the encoding and modulation functions and does not require bandwidth expansion [1]. These codes are called Ungerboeck codes or trellis coded modulation (TCM). Nearly every telephone line modem or DSL modem on the market today operating above 9.6 kb/s uses TCM.



Other coding techniques exist that are worth examining, such as hybrid ARQ. Hybrid ARQ schemes combine error detection and FEC to make more efficient use of the channel. At the receiver, the decoder first attempts to correct any errors present in the received code word. If it cannot correct all the errors, it requests retransmission using an ARQ technique.

Another technique to consider is interleaving. Many real-world channels are burst-error channels. For example, the mobile radio channel is a burst-error channel as a consequence of multipath fading. The most popular way to correct burst errors is to take a code that works well on random errors (e.g., a convolu-

tional code) and interleave the bursts to "spread out" the errors so that they appear random to the decoder. There are two types of interleavers commonly in use today, block interleavers and convolutional interleavers. Figure 1 on page 38 illustrates the operation of a block interleaver.

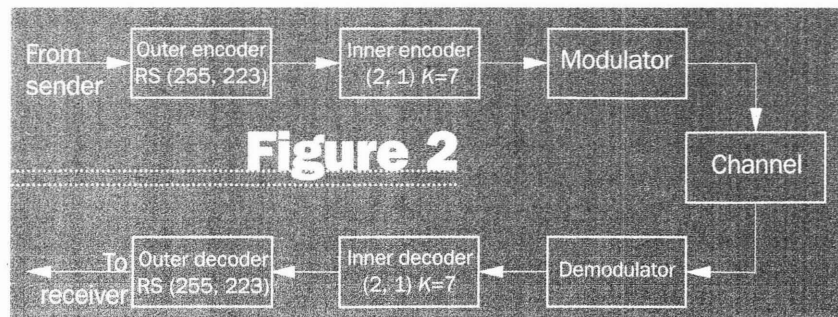
The block interleaver is loaded row by row with L code words, each with a length of n bits. These L code words are then transmitted column by column until the interleaver is emptied. Then the interleaver is loaded again, and the cycle repeats. At the receiver, the code words are deinterleaved before they are decoded. A burst of length L bits or less will cause no more than 1 bit error in any single code word. The random error decoder is much more likely to correct this

single error than the entire burst.

The parameter L is called the interleaver degree, or interleaver depth. The interleaver depth is chosen based on worst-case channel conditions. It must be large enough so that the interleaved code can handle the longest error bursts expected on the channel. The main drawback of interleavers is the delay introduced with each row-by-row fill of the interleaver. The delay is a function of the interleaver depth, which, in turn, is a function of the fade duration on the channel. The delay on some channels

can be several seconds long. This long delay is often unacceptable. On voice circuits, for example, interleaver delays confuse the unfamiliar listener by introducing long pauses between speaker transitions. Even short delays of less than one-half second are sufficient to disrupt normal conversation.

In theory, interleaving is a poor way to handle burst errors. Why? From a strict probabilistic sense, we are converting "good" errors into "bad" errors. Burst errors have structure, and that structure can be exploited (in theory).



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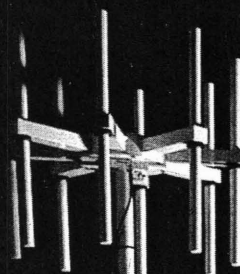


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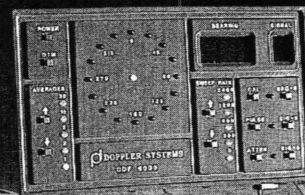


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Interleavers "randomize" the errors and destroy the structure. Despite this theoretical disadvantage, interleaving is one of the best burst-error-correcting techniques in practice. In fact, the greatest advance in coding theory in the past 15 years, turbo coding, employs a very long random interleaver [2]. Until the coding theorists discover a better way, interleaving will be an essential error-control-coding technique for bursty channels.

Yet another technique worth exploring is concatenation. When two codes are used in series, the combination is called a concatenated code. Concatenated codes often are used when a single code cannot correct all types of errors encountered on the channel. The operation of concatenated codes is best illustrated by example. Figure 2 on page 40 shows the Consultative Committee

for Space Data Systems (CCSDS) Blue Book standard for Telemetry Channel Coding (interleaving is omitted for clarity) [3].

The inner code—a rate $1/2$, $K=7$, convolutional code with Viterbi decoding—corrects most random errors and the outer code—a 16-symbol error correcting Reed-Solomon code—cleans any burst errors that slip through the Viterbi decoder. The Reed-Solomon code operates on 8-bit symbols and therefore is a very powerful burst-error correcting code. The overall code rate is simply the product of the two individual code rates, i.e., $(1/2)(223/255) \approx 0.44$. It should be noted that coding gains for concatenated codes are very large.

Next month: Error-control coding in APCO 25 Radios. ■

Jay Jacobsmeyer is president of Pericle

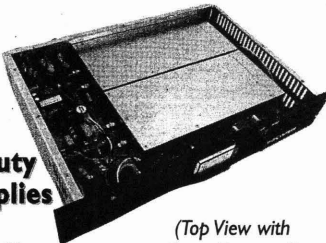
Communications Co., a consulting engineering firm located in Colorado Springs, Colo. He holds bachelor's and master's degrees in Electrical Engineering from Virginia Tech and Cornell University, respectively, and has more than 20 years experience as a radio frequency engineer.

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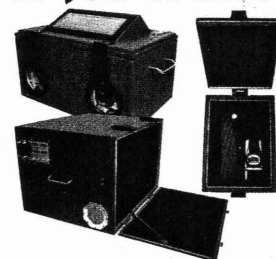
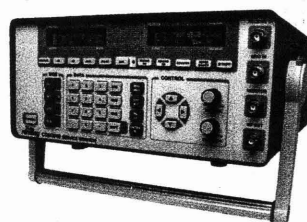
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What error-control coding can do

However, the coding's limitations also must be realized

Part 2 of 4

In the November 2005 *MRT*, we defined terms and introduced basic error-control coding concepts. This month we will cover some of the applications and limitations of codes and the theory behind their operation.

The traditional role for error-control coding was to make a troublesome channel acceptable by lowering the frequency of error events. The error events could be bit errors, message errors or undetected errors. While reducing the occurrence of undetected errors was one of the first uses of error-control coding, today's error-detection codes are so effective that the occurrence of undetected errors is, for all practical purposes, eliminated. Today, the role of error-control coding has expanded to include many new applications, including:

■ **Reduce the cost of communications systems:** Transmitter power is expensive, especially on satellite transponders. Coding can reduce the satellite's power needs because messages received at close to the thermal noise level can still be recovered correctly.

■ **Eliminate interference:** As the electromagnetic spectrum becomes more crowded with man-made signals, error-control coding will mitigate the effects of unintentional interference.

Despite these new uses of error-control coding, there are limits to what coding can do. On the Gaussian noise channel, for example, Shannon's capacity formula sets a lower limit on the signal-to-noise ratio that we must achieve to maintain reliable communications. For strictly power-limited (unlimited bandwidth) channels, Shannon's lower threshold can be expressed by E_b/N_0 equals 0.69, or -1.6 decibels (dB) [1]. In other words, we must maintain an E_b/N_0 of at least -1.6 dB to ensure reli-

able communications, no matter how powerful an error-control code we use.

Many channels, like the land mobile radio channel, are bandwidth limited. For bandwidth-limited channels with Gaussian noise, Shannon's capacity formula can be written as Equation 1 where r is the spectral bit rate in bits/s/Hz [2].

For example, consider a bandwidth-limited channel operating with uncoded quadrature phase shift keying (2 bits/symbol and a maximum

channels, but they are impractical for two-way voice channels.

A full understanding of the structure and performance of error-control codes requires a foundation in modern algebra and probability theory, which is beyond the scope of this column. Instead, I'll appeal to your intuition and common sense. Let's begin by showing how the encoder and decoder work for binary block codes.

Equation 1

$$\frac{E_b}{N_0} \geq \frac{2^r - 1}{r}$$

spectral bit rate of $r=2$ bit/s/Hz) and a required bit error rate (BER) of 10^{-5} . Without coding, this communications system requires an E_b/N_0 of 9.6 dB [2]. Shannon's formula says that to maintain reliable communications at an arbitrarily low BER, we must maintain (for $r=2$ bits/s/Hz) an E_b/N_0 of at least 1.5, or 1.8 dB.

Therefore, if we need to lower the required E_b/N_0 by more than 7.8 dB, coding can't do it. We must resort to other measures, such as increasing transmitter power. In practice, the situation is worse because no practical code achieves Shannon's lower threshold. Until recently, a more realistic coding gain for this example was 3 dB rather than 7.8 dB. The invention of turbo codes in 1993 allows us to nearly achieve Shannon's coding limit, but at the price of delay [3]. Consequently, turbo codes are good choices for data

The block encoder takes a block of k bits and replaces it with an n -bit codeword (n is bigger than k). For a binary code, there are 2^k possible codewords in the codebook. The channel introduces errors and the received word can be any one of 2^n n -bit words of which only 2^k are valid codewords. The job of the decoder is to find the codeword that is closest to the received n -bit word. How a practical decoder does this is more than we can cover here, but our examples will use a brute force look-up table method. The decoding spheres represented in Figure 1 on page 35 will be used to illustrate the decoding process.

In Figure 1, each valid codeword is represented by a point surrounded by a sphere of radius t , where t is the number of errors that the code can correct [2]. Note that codewords A and B of Figure 1 are separated by a distance d_{min} , called the minimum distance of the code. The

minimum distance of a code is defined as the smallest number of places that any two codewords in the codebook differ. Usually, codes with large minimum distance are preferred because they can detect and correct more errors.

Let's first consider a decoder that can only detect errors, not correct them.

■ **Error detection only:** The minimum distance of a code is a measure of its error-detection capability. An error-control code can be used to detect all patterns of u errors in any codeword as long as $d_{min} = u + 1$. The code also may detect many error patterns with more than u errors, but it is guaranteed to detect all patterns of u errors or less. We'll

assume that the error-detection decoder comprises a look-up table with all 2^k valid codewords stored. When an n -bit word is received by the decoder, it checks the look-up table, and if this word is one of the allowable codewords, it flags the n -bit word as error-free and sends the corresponding information bits to the user.

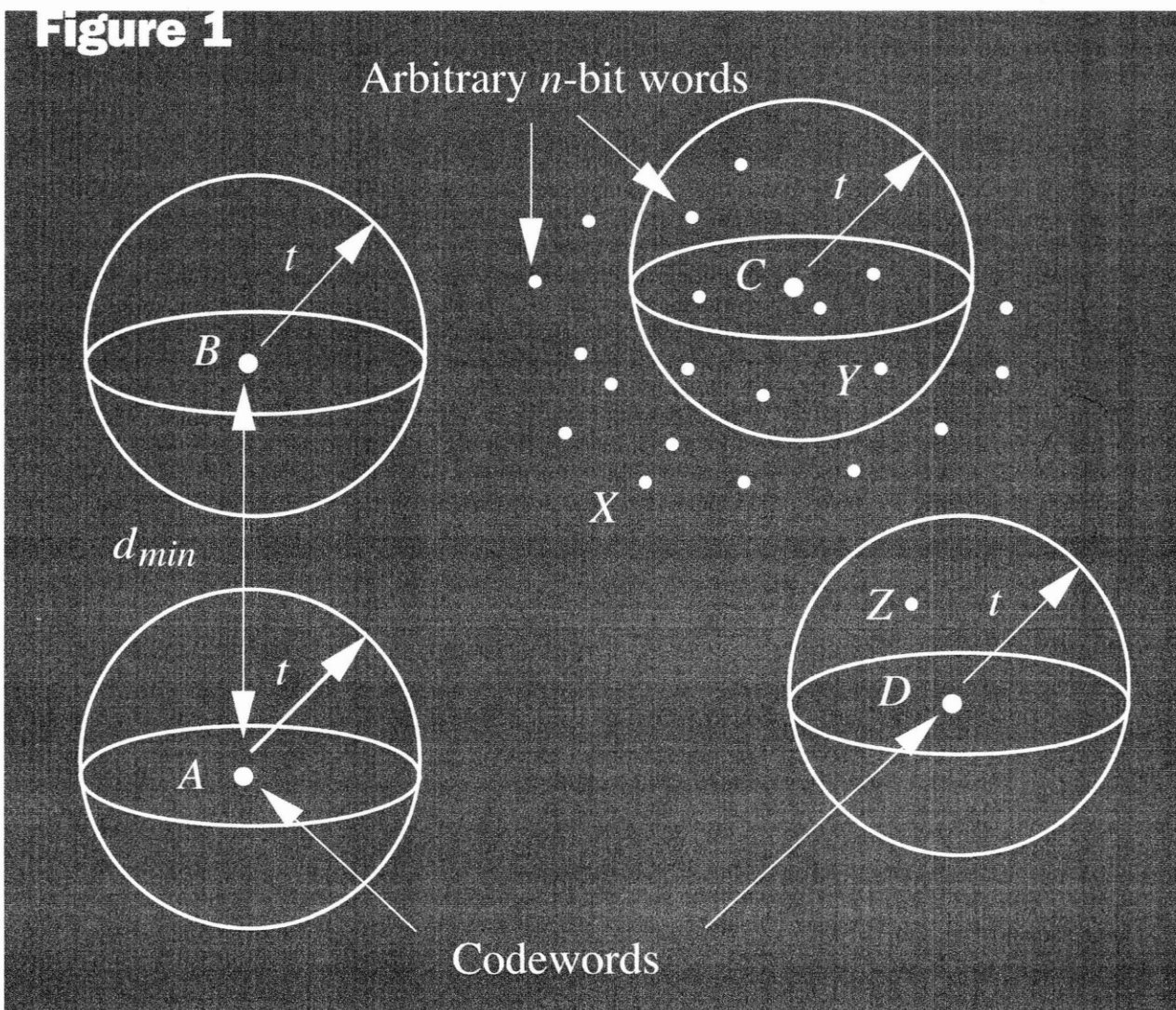
We'll use Figure 1 to illustrate three cases: no errors, a detectable error pattern, and an undetectable error pattern.

■ **No errors:** Assume the encoder sends codeword C , and the channel introduces no errors. Then codeword C also will be received, the decoder will find it in the look-up table and decoding will be successful.

■ **Detectable error pattern:** This time we send codeword C , and the channel introduces errors such that the n -bit word Y is received. Because Y is not a valid codeword, the decoder will not find it in the table and will therefore flag the received n -bit word as an errored codeword. The decoder does not necessarily know the number or location of the errors, but that's OK because we only asked the decoder to detect errors. Because the decoder properly detected an errored codeword, decoding is successful.

■ **Undetectable error pattern:** We send codeword C for the third time, and

Continued on page 51



Continued from page 35

Figure 2

0 → 00000
1 → 11111

this time the channel introduces the unlikely (but certainly possible) error pattern that converts codeword C into codeword D . The decoder can't know that codeword C was sent and must assume that codeword D was sent instead. Because codeword D is a valid codeword, the decoder declares the received n -bit word error-free and passes the corresponding information bits on to the user. This is an example of decoder failure.

Naturally, we want the decoder to fail rarely, so we choose codes that have a small probability of undetected error. One of the most popular error-detection codes is the shortened Hamming code, also known as the cyclic redundancy check (CRC).

Comparing the spheres surrounding codewords A and B in Figure 1, we see that the error-correcting capability of a code is given by $d_{\min}=2t+1$ (this is the minimum separation that prevents overlapping spheres). In other words, a code with $d_{\min}=3$ can correct all patterns of 1 error, one with $d_{\min}=5$ can correct all patterns of 2 errors, and so on. A code can be used to correct t errors and detect v additional errors as long as $d_{\min} \geq 2t + v + 1$. Now refer to Figure 1 and consider the following error-decoding cases: correct decoding, decoding failure and error detection without correction.

■ **Correct decoding:** Assume that codeword C is sent, and the n -bit word Y is received. Because Y is inside C 's sphere, the decoder will correct all errors.

■ **Decoding failure:** This time we send codeword C , and the channel gives us n -bit word Z . The decoder has no way

of knowing that codeword C was sent and must decode to D since Z is in D 's sphere. This is an example of error-correction decoder failure.

■ **Error detection without correction:** This case shows one way that an error-correction code can be used to also detect errors. We send codeword C and receive n -bit word X . Since X is not inside any sphere, we won't try to correct it. We do, however, recognize that it is an errored codeword and report this information to the user.

Table 1

**Post-Decoding
Probability of
Bit Error**
(5,1) Repetition Code

Input BER	Output BER
10^{-2}	9.9×10^{-6}
10^{-3}	1.0×10^{-8}
10^{-4}	1.0×10^{-11}
10^{-5}	1.0×10^{-14}

In the last example, we could try to correct n -bit word X to the nearest valid codeword, even though X was not inside any codeword's sphere. A decoder that attempts to correct all received n -bit words regardless of whether they are in a decoding sphere is called a complete decoder. On the other hand, a decoder that attempts to correct only

n -bit words that lie inside a decoding sphere is called an incomplete, or bounded-distance decoder. Bounded-distance decoders are much more common than complete decoders.

Now let's apply what we've learned to a simple error correction code, known as the repetition code. Consider a (5, 1) repetition code that repeats each bit four times. Figure 2 depicts such an encoder.

The decoder takes 5 bits at a time and counts the number of 1s. If there are three or more, the decoder selects 1 for the decoded bit. Otherwise, the decoder selects 0. The minimum distance of this code is 5, so it can correct all patterns of two errors. To compute the error performance of this code, consider a random error channel with a probability of bit error of p . After decoding, the probability of bit error is simply the probability of three or more bit errors in a 5-bit codeword. This probability is computed for several values of p with results listed in Table 1.

Next month: Error-control coding techniques. ■

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- [1] C. E. Shannon, "A Mathematical Theory of Communication," Bell System Technical Journal, vol. 27, pp. 379-423, 1948.
- [2] R. E. Blahut, Theory and practice of error control codes, Reading, Massachusetts: Addison-Wesley Publishing Co., 1983.
- [3] C. Berrou, A. Glavieux and P. Thitimajshima, "Near Shannon limit error-correcting coding and decoding: Turbo-codes," IEEE Transactions on Communications, pp. 1261-71, October 1996.

The next level of intermodulation

Here's what to do when dealing with amplifier signals that are unequal

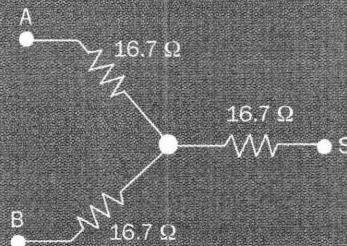
In the April edition, we lightly touched on the subject of intermodulation (I-M) and small-signal radio frequency amplifiers. We discussed how two equal-level tones (RF signals), A and B, when applied to the amplifier input, are used to produce third-order intermodulation products at the output of the amplifier. We left the reader hanging, somewhat, with the question: "What if tones A and B are not equal?" In this article, we provide the answer.

First, a little background about interconnecting the test equipment and the device under test (DUT) is in order. Two signal generators labeled A and B are used to generate the two continuous wave tones, A and B. The outputs of the two signal generators are combined and fed to the input of the DUT (a small-signal RF amplifier in this case), and the frequency and output level of the two generators are adjusted to produce third-order I-M products at the output of the RF amplifier. The test setup is shown in Figure 1.

A hybrid combiner is preferred over a simple resistive combiner. A resistive combiner like the one shown in Figure 2, while providing proper impedance matching, offers little isolation between signal generators. A high degree of isolation between combiner ports A and B is necessary to prevent interaction between the signal generators. Don't even think about using a simple Tee connector as a combiner for this test. A hybrid combiner is made up of an RF transformer and resistor network. The isolation between ports A and B is achieved by cancellation of signals at opposite ports.

For example, the signal at port A arrives at port B via two separate paths. One path is through a transformer, and the other path is through a resistor. The path through the RF transformer is 180° out of phase with the signal through the resistor. With the signals being equal—or nearly so—cancellation occurs. The same is true for the signals traveling from port B to port A. Referring to Fig-

Figure 2



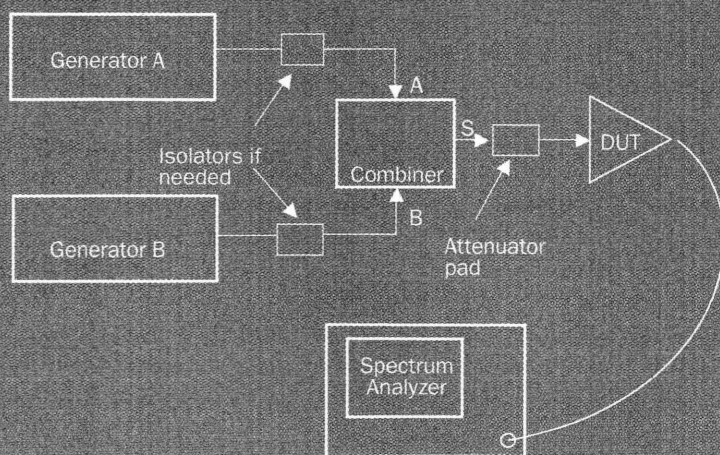
A resistive combiner such as this could be used to combine the signals from two signal generators. However, not enough isolation would be provided between the signal generators. Additional means of isolation would be required.

ure 1, RF isolators can be used with hybrid combiners for additional isolation. But they must be used with resistive combiners in order to achieve the required degree of isolation.

Another important point to note is that having a good 50 ohm load at the sum (S) port is necessary to optimize the isolation between ports A and B. A mismatch at the S port will adversely affect the isolation between ports A and B, possibly resulting in undesirable interaction between the signal generators. An attenuator pad connected as shown can improve the isolation figure.

In order to check for interaction between signal generators, vary the output level of generators A and B by ± 10 dB and observe the effect on the opposite signal on the spectrum analyzer. For example, increase and decrease the level of generator A by 10 dB while closely monitoring the level of signal B on the spectrum analyzer. If the level of signal B is affected by changes in the level of generator A, there is insufficient isolation between the two signal generators. Also, make sure that I-M signals

Figure 1



This test setup is used to check the intermodulation distortion (IMD) performance of the DUT (device under test), a small-signal RF amplifier. See text for full discussion.

are not being formed in the signal generators. You can check this by changing the attenuation in the pad between the DUT and the combiner. If you see a dB-for-dB change in the I-M signal on the spectrum analyzer, the I-M signal is being formed at a point prior to the attenuator pad. If you see a 3-for-1 dB change in the third-order I-M signal on the spectrum analyzer, the I-M signal is being formed at a point after the attenuator pad.

Suppose that the amplifier in Figure 1 has a third-order intercept point (TOIP) of +10 dBm. Signal generators A and B are set to 150 MHz and 152 MHz, respectively. The output levels of generators A and B are set to produce a level of -10 dBm for each tone (A and B) at the input to the DUT (amplifier). Since both of the fundamental tones are equal in amplitude, the two third-order intermodulation products also are equal in amplitude. In Figure 3, each horizontal scale division on the spectrum analyzer represents 1 MHz. Thus, the 2A-B intermodulation product is at 148 MHz, and the 2B-A intermodulation product is at 154 MHz. Notice that the levels of the I-M products are each at -50 dBm. Also note in Figure 4 that the fundamental tones, A and B, are not of equal level. This results in I-M products that are not of equal level. The frequency spacing remains unchanged.

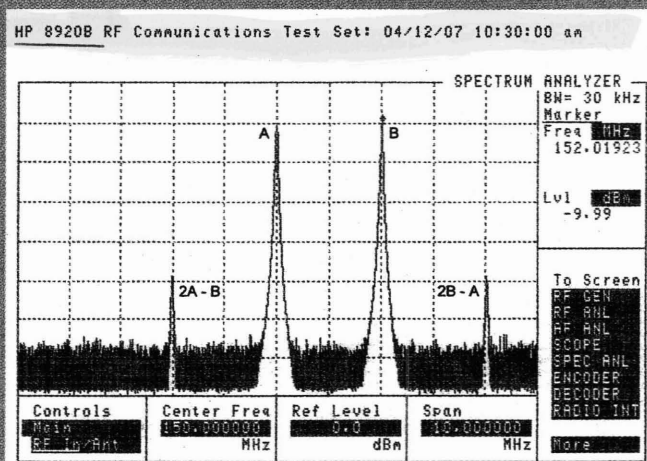
Figure 5 on page 48 is a nomogram that can be used to determine the output level of the third-order I-M products, given the amplifier's third-order intercept figure and the level of fundamental tones, A and B, at the output. First, a description of the scales of the nomogram is in order. The first scale on the far left is used to represent the level of the second-order term. The second scale is used to represent the level of the first-order term. In the expression 2A-B, A is the second-order term, and B is the first-order term. Conversely, in the expression 2B-A, B is the second-order term, and A is the first-order term. The third scale is used to adjust the TOIP of the amplifier. If the

TOIP of the amplifier is 0 dBm, no adjustment is necessary.

Now, let's do an example on the nomogram. Suppose that an amplifier has a TOIP of +25 dBm. The level of the fundamental tone A at the output is -20 dBm and the level of the fundamental tone B at the output is -30 dBm. Now, we want to determine the level of the two

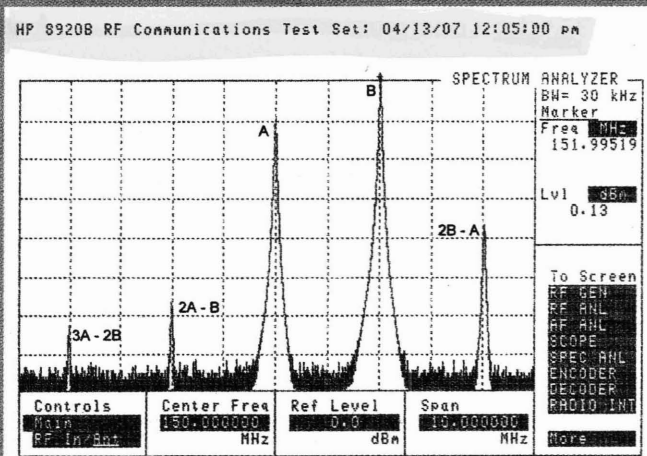
third-order I-M products at the output, 2A-B and 2B-A. First, we will determine the level of the 2A-B intermodulation signal. The blue dashed line on the nomogram in Figure 5 represents this procedure. Start on the left scale for the second-order term at -20 dBm, represented by point 1 on the blue-dashed line. Move straight over to the second scale repre-

Figure 3



This spectrum analyzer display shows the two CW tones (A and B) and the resultant third-order I-M products (2A-B and 2B-A). See text for details.

Figure 4



Here, tones A and B are unequal. Tone B is at 0 dBm and tone A is at -10 dBm. This results in the unequal I-M products as shown in the display. Notice, too, that a fifth-order I-M product appears on the left in the display. See text for full discussion.

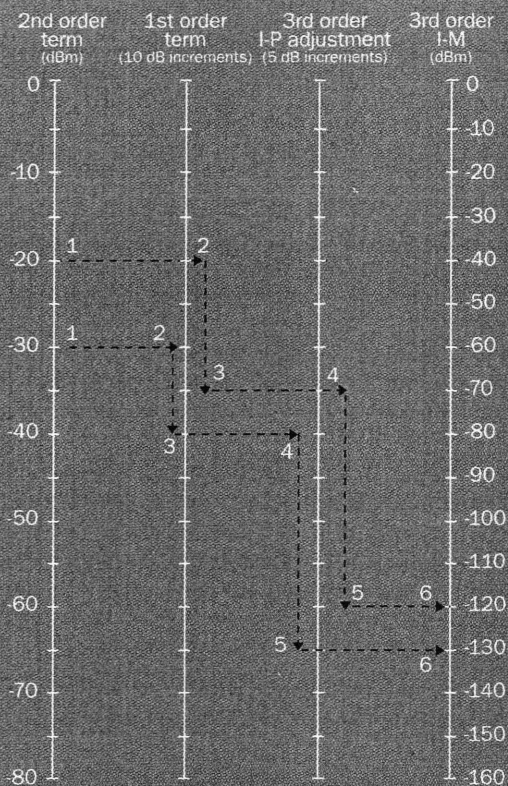
Figure 5

Figure 5: This nomogram can be used to determine the level of the third-order intermodulation product produced in the output of an amplifier, given the output level of the fundamental tones, A and B, and the third-order intercept point of the amplifier.

senting the first-order term at point 2. Since the first-order term is -30 dBm, we move downward by three increments to point 3 on the second scale. Next, move directly over to the third scale representing the third-order intercept adjustment at point 4. If the TOIP was 0 dBm, we would not need to make any adjustment. But, since the third-order intercept point is +25 dBm, we must move downward by five increments to point 5 on the third scale. Then, move directly over to the fourth scale at point 6, which represents -120 dBm for the third-order I-M level. Thus, the level of the 2A-B intermodulation product is -120 dBm. The level of the 2B-A intermodulation product is determined the same way, as depicted by the red dashed line on the nomogram.

The nomogram in Figure 5 is very intuitive, and in studying it, several things become apparent. For example, notice that a 10 dB change in the second-order term will change the third-order I-M level by 20 dB. A 10 dB change in the first-order term will change the third-order I-M level only by 10 dB, while a 10 dB change in both

terms (in the same direction) will result in a 30 dB change in the output I-M level (in that direction).

Note that on the third scale—the TOIP adjustment—each increment represents a 5 dB change and that a 5 dB change in the third-order intercept point will result in a 10 dB change in the third-order I-M level. For a given output level of the fundamental A and B tones, increasing the TOIP results in a decrease in the I-M signal level. This is why it is important to use an amplifier with a very high TOIP figure to keep the resultant I-M signal levels low. The formulas for calculating the third-order I-M levels can be found in Equation 1.

It is important to note here that this discussion centers on the output TOIP and uses the tone levels, A and B, at the output of the amplifier. Thus, the figure listed is the output TOIP. The input TOIP would be less than the output TOIP by an amount equal to the gain of the amplifier. (Typically, RF amplifiers are rated in terms of the output TOIP figure.)

It is important to remember, too, that the TOIP point is an extrapolated figure based on the point at which the third-order I-M product is equal to the level of one of the two equal fundamental tones, A or B. This point occurs at a point 15 dB (plus/minus a few decibels) above the 1 dB compression point. Since it is well beyond the point of compression, the TOIP can't be reached in a practical amplifier. Still, the figure is useful in determining low signal-level performance of an RF amplifier.

Until next time—*stay tuned!*

Equation 1

$$P_{(2A-B)} = 2P_A + P_B - 2P_{I3}$$

$$P_{(2B-A)} = 2P_B + P_A - 2P_{I3}$$

where: $P_{(2A-B)}$ and $P_{(2B-A)}$ = power level of I-M signal in dBm
 P_A and P_B = power level of signals A and B, respectively, in dBm
 P_{I3} = power level of TOIP in dBm.

An introduction to 4.9 GHz

Before utilizing the band, a little background might be helpful.

Starting in 2003, the federal government has made available 50 MHz of new radio spectrum between 4.940 MHz and 4.990 MHz for public-safety use. This spectrum is designated for fixed and mobile broadband wireless services. Communications in this band must support the protection of life, health or property, with proposed uses including wireless local area networks, or WLANs; mobile data; video security; voice over IP, or VoIP; and ad hoc networks.

Although permanent point-to-point fixed installations are allowed with certain restrictions, the main purpose of the band is for temporary point-to-point links and mobile operations, which take priority over permanent fixed installations. This article provides a basic primer for navigating the 4.9 GHz airwaves.

Prior to 2003, the 4.9 GHz band (4.940 GHz to 4.990 GHz) in the U.S. was allocated to federal government fixed and mobile services. The band was used for fixed services such as conventional point-to-point microwave, tactical radio relay, high-power tropospheric scatter systems, and for mobile services such as control of remote piloted vehicles, video and data telemetry links, target drone control links, fleet defense systems and tethered aerostat systems.

In 1999, the 4.9 GHz band was transferred from federal government use to non-government use in accordance with the provisions of the Omnibus Budget Reconciliation Act. In 2000, the FCC released a Notice of Proposed Rulemaking that suggested allocating the 4.9 GHz band to non-government fixed and mobile services, excluding aeronautical mobile service, on

a co-primary basis, and allowing for flexible use of the band. The FCC also tentatively decided not to designate the band exclusively for public-safety use. However, many local government entities and national public-safety associations argued that public safety urgently needed this spectrum for broadband wireless services. In its second Report and Order (R&O), the commission reversed itself and reserved the 4.9 GHz band exclusively for public safety [1].

The FCC issued a third R&O in May 2003 that defined additional rules for eligibility and use of the 4.9 GHz band, and for the first time allowed public-

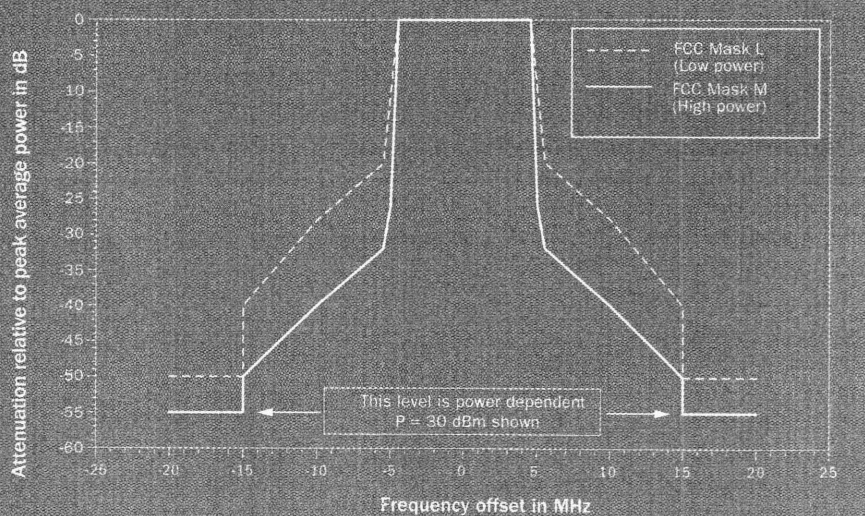
license; Part 90.1213 defines the channelization of the band; Part 90.1215 defines the power limits for radios operating in the band; and Part 90.210 defines the emission masks for the band.

The FCC allows channels to be aggregated to channel bandwidths of 1, 5, 10, 15, or 20 MHz. The maximum channel size is 20 MHz. FCC channels are listed in Table 1.

An emission mask defines how much spectrum the signal may occupy. Today, the FCC authorizes two masks for use in the 4.9 GHz band: Emission mask *L* for low-power

Figure 1

FCC 10 MHz emission masks
(Masks scale proportionately for 5 MHz and 20 MHz channels)



safety agencies to apply for and receive licenses to operate in the band [2]. FCC rules governing the 4.9 GHz band are found in Part 90, Title 47 of the Code of Federal Regulations [3]. Sections of Part 90 relevant to the 4.9 GHz band include the following: Part 90.523 defines who is eligible to hold a 4.9 GHz

devices and emission mask *M* for high-power devices. The *M* mask provides better adjacent channel protection and was adopted first.

The *L* mask is nearly identical to the mask defined in the IEEE 802.11 standards. It was chosen to let public-safety agencies and equipment vendors ex-

exploit the economies of scale created by existing 5 GHz commercial off-the-shelf devices, thereby reducing costs and time to market. Both emission masks are plotted for 10 MHz in Figure 1 on the preceding page.

operating with a 13 dBi antenna also has an EIRP of 40 dBm—both systems will result in the same receive power if the receiving station is in the main lobe of the transmitting antenna.

Both low- and high-power devices

directional antennas with gain up to 26 dBi at maximum authorized power. However, directional antenna gain may exceed 26 dBi if both transmitter power and power spectral density are reduced decibel-for-decibel by the amount the directional antenna gain exceeds 26 dBi.

EIRP limits for directional antenna systems are summarized in Table 3 on page 38.

The emission mask was a contentious issue in 2003 and 2004. Emission mask *M*—also known as the “tight” mask—originally was the sole emission mask authorized. But neither IEEE 802.11a or 802.11j radios could meet this emission mask, so there was no industry-standard 4.9 MHz product available to public-safety agencies.

Consequently, the National Public Safety Telecommunications Council (NPSTC) and other agencies filed comments with the FCC objecting to the original emissions mask on the grounds it would preclude the use of industry-standard 802.11 radios and would result in expensive, proprietary

Higher-power devices will of course extend the range and reliability of 4.9 GHz networks, so from that perspective, emission mask *M* is preferred.

Emission mask, transmitter power and effective isotropic radiated power (EIRP) are tightly coupled in the FCC rules. Part 90.1215 creates a relatively complicated set of power limitations that we will summarize here.

Part 90.1215 limits both transmitter power and EIRP. The power limits for low- and high-power devices are listed in Table 2. Note that within each category, the power density (dBm/MHz) is the same, regardless of channel bandwidth. The power-density limit for low-power devices is 7 dBm /MHz, while the power-density limit for high-power devices is 20 dBm/MHz.

EIRP is the product of transmitter power and antenna gain (relative to isotropic radiated power). Assuming that the receiving station is operating in the main lobe of the transmitting antenna, EIRP determines the power received, not the transmitter power. For example, a 30 dBm transmitter operating with a 10 dBi antenna has an EIRP of 40 dBm, but a 27 dBm transmitter

may use omnidirectional or directional antennas with gains up to 9 dBi at maximum transmitter power.

Low-power devices may use directional antennas with gain greater than 9 dBi if both transmitter power and power spectral density are reduced

Table 1

FCC channels in the 4.9 GHz band

Channels	Center frequency (MHz)	Bandwidth (MHz)
1-5	$4940.5 + (N-1) \times 1 \text{ MHz}$	1
6-13	$4947.5 + (N-6) \times 5 \text{ MHz}$	5
14-18	$4985.5 + (N-14) \times 1 \text{ MHz}$	1

Table 2

FCC power limits at 4.9 GHz

Channel bandwidth	Low-power device (FCC Mask L)	High-power device (FCC Mask M)
1 MHz	7 dBm	20 dBm
5 MHz	14 dBm	27 dBm
10 MHz	17 dBm	30 dBm
15 MHz	18.8 dBm	31.8 dBm
20 MHz	20 dBm	33 dBm

decibel-for-decibel by the amount the directional antenna gain exceeds 9 dBi.

High-power devices used for point-to-point or point-to-multipoint operation (fixed or temporary) may use

devices that would stifle innovation. The FCC relented and eventually adopted two masks for 4.9 GHz, the original *M* mask for high-power devices and the *L* mask for low-power

devices, in November 2004 [3].

Thus, the prohibition on the 802.11a mask has been lifted, but the power limitations of the "loose" L mask create an incentive to use tight mask devices.

At the time of this writing, some manufacturers are developing or have fielded devices that are 802.11 standard-compliant and meet the tight mask, so the original problem may soon be overcome. ■

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References:

[1] Second Report and Order and Further Notice of Proposed Rule Making, "In the Matter of the 4.9 GHz Band Transferred from Federal

Table 3

FCC EIRP limits at 4.9 GHz (directional antennas)

Channel bandwidth	Low-power device (FCC Mask L)	High-power device (FCC Mask M)
1 MHz	16 dBm	46 dBm
5 MHz	23 dBm	53 dBm
10 MHz	26 dBm	56 dBm
15 MHz	27.8 dBm	57.8 dBm
20 MHz	29 dBm	59 dBm

Government Use," WT Docket 00-32, Adopted February 14, 2002.

[2] Memorandum Opinion and Order & Third Report and Order, "In the Matter of the 4.9 GHz Band Transferred from Federal Government Use," WT Docket 00-32, Adopted April 23, 2003.

[3] Title 47 of the U.S. Code of Federal Regu-

lations, Part 15, Part 90, October 1, 2005.

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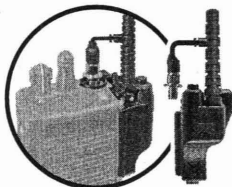
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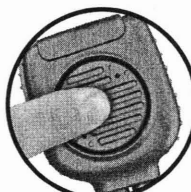
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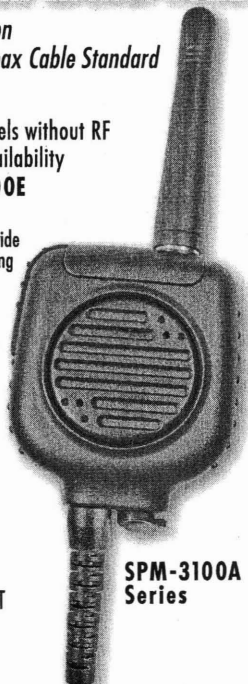
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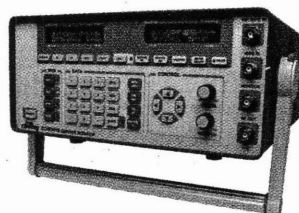


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TECHNICALLY SPEAKING

SWR, impedance and line loss

Directional wattmeter proves invaluable

Standing waves on a transmission line affects both the impedance seen at the input and the overall loss of the transmission line — the higher the SWR, the greater the effect on these parameters.

This column deals with the practical aspects of SWR, line loss and input impedance.

Forward/reflected power

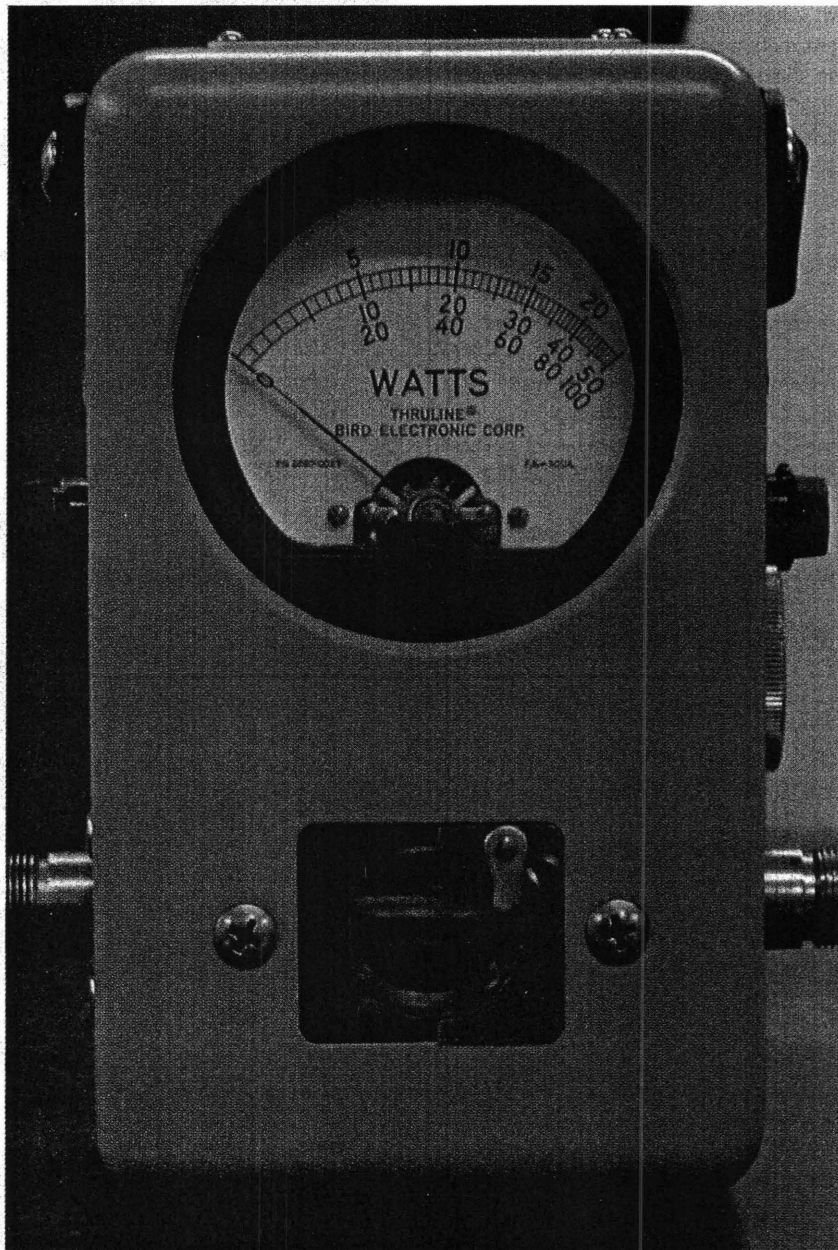
In last month's column, we discussed the use of the directional wattmeter in making measurements of forward and reflected power on the transmission line.

Figure 1 shows the setup in which the directional wattmeter is used to measure forward and reflected power. At A., the directional wattmeter is set to measure forward or incident power.

Suppose that the transmission line has a theoretical loss of 0dB. This means that all of the forward or incident power at the input of the line will reach the antenna. If the antenna is perfectly matched to the line, all of the forward power will be radiated by the antenna and no power will be reflected back down the line. If the transmitter output power is 100W, the power reaching the antenna is 100W and the reflected power is zero.

Next, suppose that the antenna is not perfectly matched to the transmission line impedance. In this case, some of the power will be reflected back down the line toward the source or input. Since the line has no loss, all of the reflected power will reach the transmitter.

Suppose that the load impedance is 150 ohms purely resistive and the transmission line is a length such that the impedance at the input of the line is $30 - j40\Omega$. This would



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Insertion type RF wattmeter for measurement of power flow and load match in 50-ohm coaxial transmission lines.

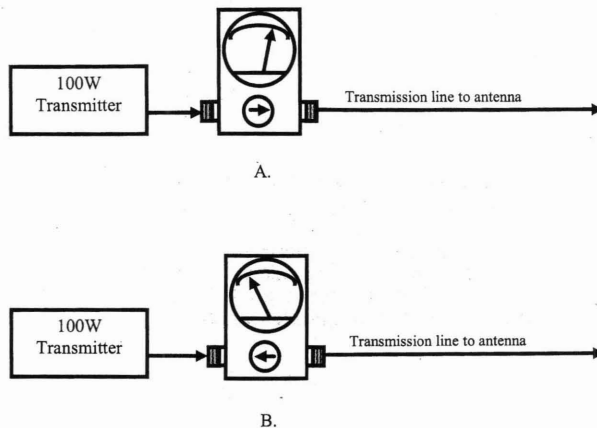


Figure 1: At A., the directional wattmeter is set to read the forward or incident power at the input to the transmission line. At B., the wattmeter is set to read the reflected power at the input to the transmission line.

cause an SWR of 3:1. Since the line is lossless, the SWR would be the same at all points on the line. An SWR of 3:1 means that the reflected power would be 25 percent of the incident or forward power.

See Figure 2. The length of the jumper cable is such that the length of the cable plus the through-line section of the wattmeter is

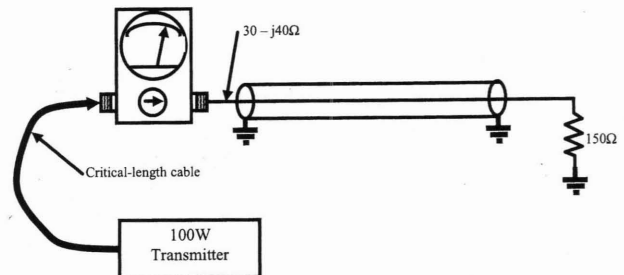


Figure 2: Here, the transmission line is terminated in a load impedance of 150Ω. The line length is such that the impedance at the input of the line is $30 - j40\Omega$. The jumper cable between the wattmeter is such that the length of the cable plus the through-line section of the wattmeter is equal to one-half wavelength at the operating frequency.

equal to one-half wavelength at the operating frequency. When calculating this length, be sure to consider the cable's velocity factor. If the transmitter output power was 100W into a perfectly matched load impedance, then the transmitter's output circuit should be adjusted to provide a net (forward minus reflected) power of 100W by tuning the transmitter output for a conjugate match. The conjugate match impedance in this case would be $30 + j40\Omega$, since the line input impedance is $30 - j40\Omega$.

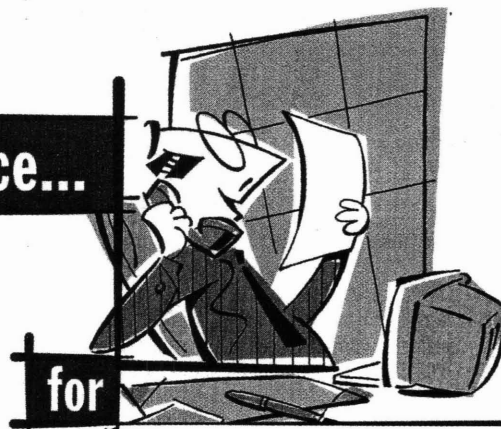
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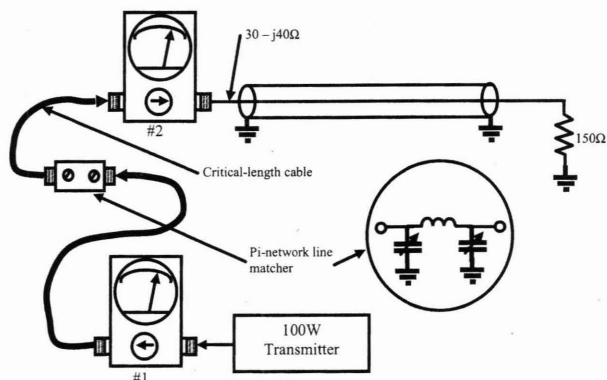


Figure 3: Here, a pi-network line matcher is inserted between the transmitter and the line input. The line matcher is tuned for minimum reflected power at the transmitter and maximum forward power into the line. If two directional wattmeters are connected as shown here, when the line matcher is properly adjusted wattmeter #1 will indicate 100W forward power and 0W reflected power. Wattmeter #2 will indicate 133W forward power and 33.3W reflected power. The reflected power is still there but the conjugate match provides additional power to overcome the loss of the reflected power. The inset shows the makeup of the line matcher. It is just a pi-network with tunable input/output capacitors.

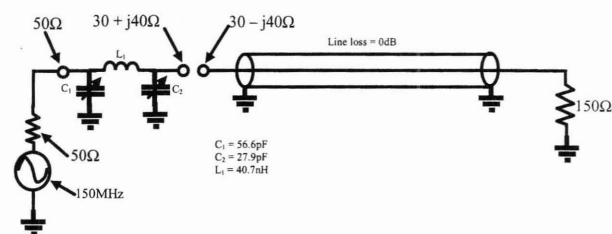


Figure 4: The reflected power traveling back down the line sees an impedance of $30 + j40\Omega$ looking into the output side of the line matcher. This provides a conjugate match for the $30 - j40\Omega$ impedance at the input side of the line. This re-reflects the reflected power to add to the forward power so that any loss from the reflected power is overcome by the reflection gain from the conjugate match.

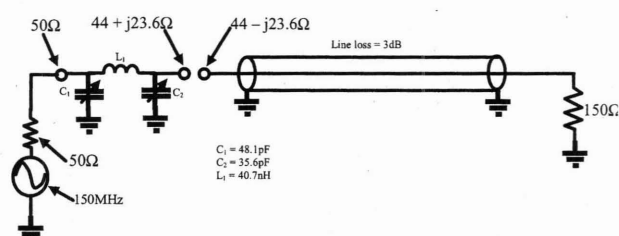


Figure 5: Here, the transmission line is the same length as in Figure 4 and the load impedance is the same, 150W. However, the line loss in this example is 3dB. This changes the impedance looking into the input of the line. The impedance looking into the line input is now $44 - j23.6\Omega$. The pi-network line matcher must be adjusted to provide a conjugate match to the new impedance. The new component values are listed in the figure.

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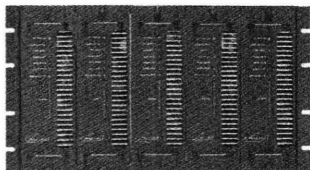
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If the transmitter output is capable of being tuned to provide the conjugate match, the forward power should read 133.3 watts and the reflected power should read 33.3 watts. The reflected power (still 25 percent) adds to the forward power of 100W to produce the total incident or forward power of 133.3 watts. If the transmitter output can't be adjusted to compensate for the mismatch, try changing the length of the line between the line input and transmitter output. This will change the impedance seen by the transmitter.

Several different lengths should be tried until a length is found to provide an impedance into which the transmitter can be tuned. Don't change the length of the jumper cable between the wattmeter and the transmitter. It must remain the same so that the situation doesn't change when the wattmeter and jumper cable are removed from the line.

If the transmitter isn't cable of being tuned into the mismatched impedance, try using a device called a line matcher or Z-matcher between the line input and the transmitter output. See Figure 3. With the pi-network line matcher in the line and properly tuned, wattmeter No. 1 will read 100W forward power and 0W reflected power. Wattmeter No. 2 will indicate 133.3W forward power and 33.3W reflected power. The net power is 133.3W minus 33.3W or 100W.

Therefore, with the conjugate match provided by the line matcher, the reflected power isn't lost. It is simply redirected back

down the line to the load. The power actually dissipated in the load or radiated by the antenna will be the same as if there were no reflected power. This no longer holds true when the line has significant matched-line loss and a high SWR.

Figure 4 shows the actual parts values for the pi-network line matcher for a frequency of 150MHz to match the line impedance of $30 - j40\Omega$ to the 50-ohm source impedance.

Line loss

In Figure 5, the line has been replaced with a line that is the same electrical length but which has a loss of 3dB. Notice that the line input impedance has changed from $30 - j40\Omega$ to $44 - j23.6\Omega$. The pi-network line matcher is retuned to again provide a conjugate match to the line input. See details in Figure 5. However, this is not the only complication caused by the line loss.

Refer to Figure 6. Assume that the line matching pi-network has been adjusted for optimum conjugate match to the transmission line. First, assume that initially 100W is applied to the input of the transmission line. Since the line has a loss of 3dB the forward power arriving at the antenna is 50 watts. Since the 150Ω load will present a SWR of 3:1, 25 percent of the forward power will be reflected. See the formula below.

$$R = 100 \left(\frac{S-1}{S+1} \right)^2$$

where R is percentage of power reflected and S is SWR.

Now, 25 percent of the forward power at the antenna, or 12.5W, is reflected from the antenna back down the line. Now, the reflected power encounters the same line loss of 3dB.

Therefore, the reflected power arriving back at the input of the line is 6.25W. By the nature of the conjugate matching network, the forward power is going to increase to an amount determined from the formula below.

$$P = \frac{P_f^2}{P_f - P_r}$$

where P is final forward power under state-state conditions, P_f is initial incident power

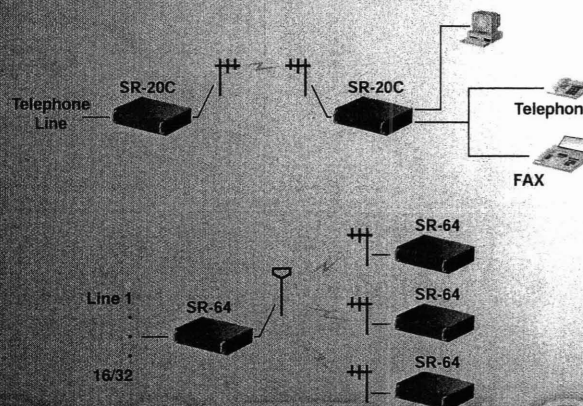
and P_r is initial reflected power.

In our example, the solution is found from the following equation.

$$P = \frac{100^2}{100 - 6.25} = \frac{10000}{93.75} = 106.67W$$

Since the reflected power at the line input was 6.25 percent we multiply the P by .0625 to get the reflected power of 106.67W x

.0625 = 6.67W. Therefore, at the line input, the steady-state forward power is 106.67W and the steady-state reflected power is 6.67W. The net power input to the line is still 100W — 106.67W minus 6.67W is 100W. Since this line has a 3dB loss, the forward power at the load is now 53.33W and the reflected power at the load is 13.34W for a net power of 53.33W minus 13.34W = 39.99W.



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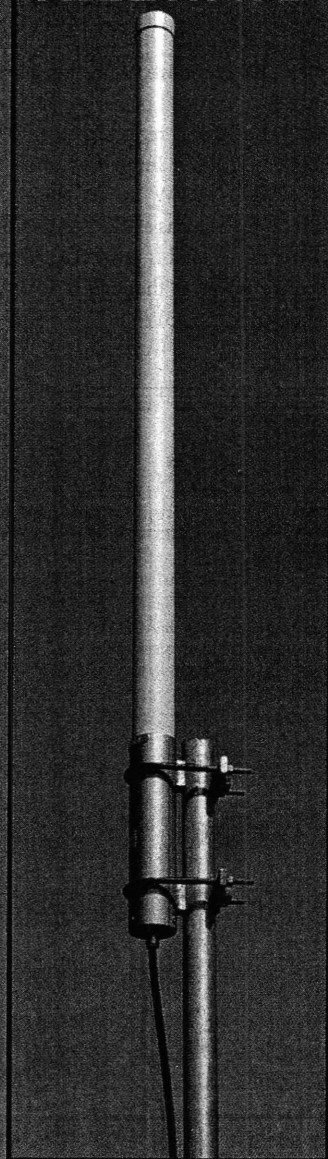


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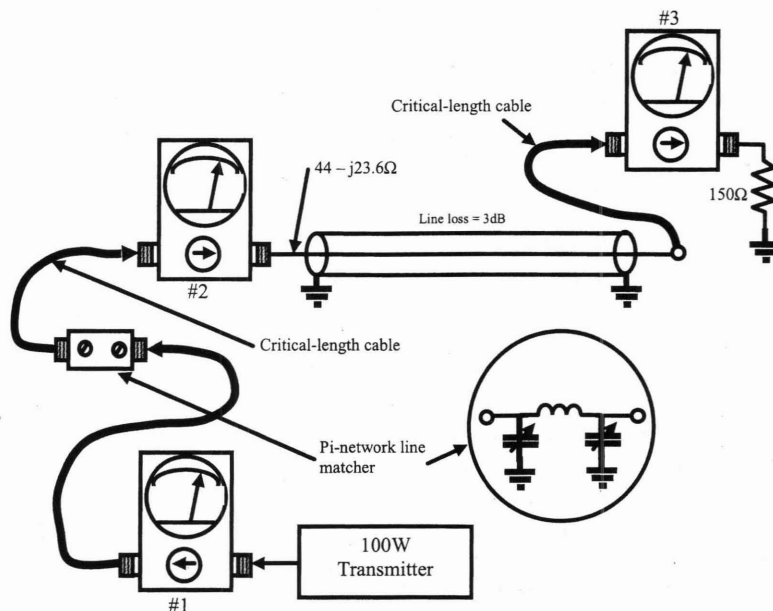


Figure 6: This is the same setup shown in Figure 3 except that the line loss is 3dB. This changes the input impedance of the line to $44 - j23.6\Omega$. When the line matcher is properly adjusted to present a conjugate match the forward power on wattmeter No. 1 will read 100W and the reflected power on wattmeter No. 1 will read 0W. What about wattmeter No. 2? What about wattmeter No. 3? See text for detailed explanation.

So, let's call it 40W at the antenna.

Remember, we had a net power of 100W at the line input and a net power of 40W at the antenna.

This represents a loss of more than 3dB. The actual loss is found from the equation below.

$$L = 10 \log \left(\frac{40}{100} \right) = 3.98 \text{ dB}$$

The matched-line loss was only 3dB so the SWR has caused an additional line loss of approximately 1dB to bring the total line loss to approximately 4dB.



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This is why many amateur radio operators prefer to use open feeder or ladder line for a transmission line. Since the matched-line loss of this type of line is quite low, the

Don't be overly concerned about SWR.
All of that reflected power isn't lost after all
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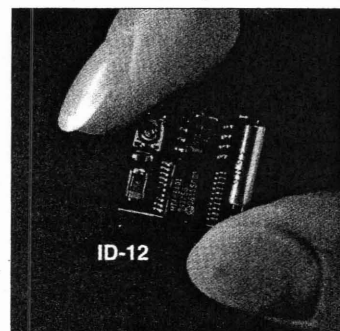
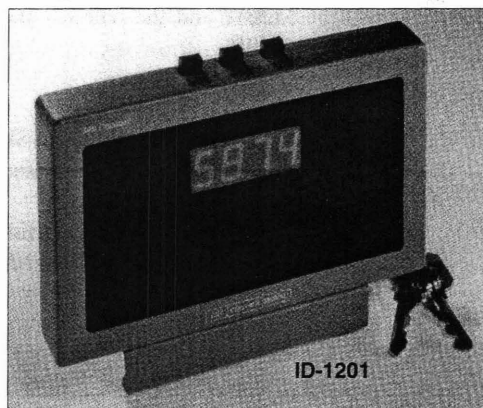


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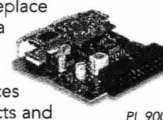
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Modern solid-state transmitters have a fold-back SWR protection circuit that will cause a significant power reduction in the presence of a high SWR at the transmitter output.

antenna can present a very high SWR without really causing much loss.

Getting back to Figure 6, assuming that the line matcher has been properly adjusted for a conjugate match, the forward power reading on wattmeter No. 1 will be 100W and the reflected power will be 0W. The forward power measured by wattmeter No. 2 will be 106.67W and the reflected power will be 6.67W. The forward power measured by wattmeter No. 3

will be 53.33W and the reflected power will be 13.34W.

Summary

Don't be overly concerned about SWR. All of that reflected power isn't lost after all — it is just recycled.

Changing the length of a transmission line will change the input impedance if the load isn't properly matched to the line. By experimentally adjusting the line length, it

is possible to find a length into which the transmitter can be properly loaded.

A line matcher can be used to provide a proper match in cases where adjusting the line length isn't practical or desirable.

Line length doesn't matter when the line is properly terminated, the input impedance will remain unchanged with varying lengths of line.

High SWR on a line with high matched-line loss will produce significant additional loss on the line.

Modern solid-state transmitters have a fold-back SWR protection circuit that will cause a significant power reduction in the presence of a high SWR at the transmitter output. Such a reduction can create a significant loss. With such a circuit, it is important to keep the SWR at the transmitter output to a very low level.

For further reading on the subject, check out Walter Maxwell's "Reflections II and The ARRL Antenna Book," 19th edition.

Until next time — stay tuned! ■

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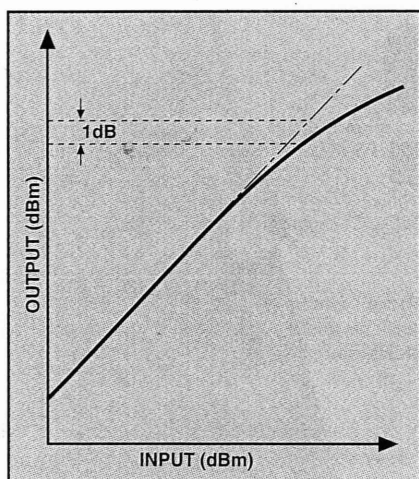


Figure 1. The 1dB compression point of an amplifier.

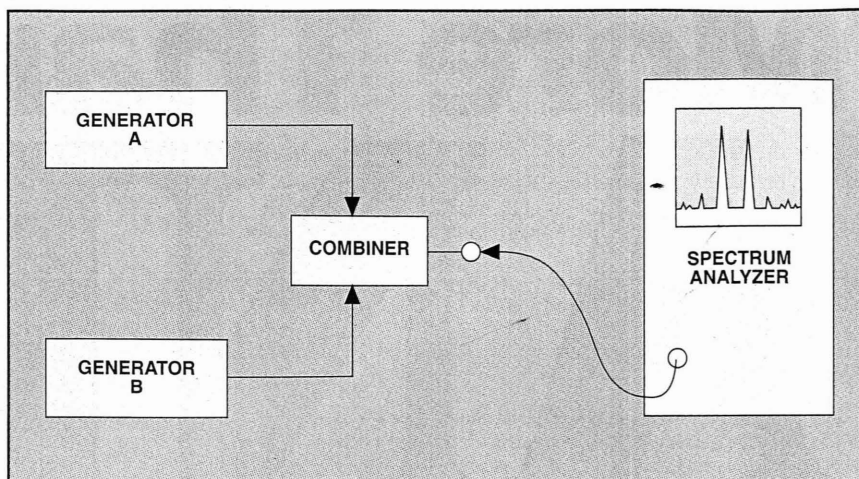


Figure 2. A simplified test setup for determining a spectrum analyzer's dynamic range.

Intermodulation distortion

By Harold Kinley, C.E.T.

An overdriven or overloaded amplifier distorts the signal it is intended to amplify. Intermodulation distortion (IMD) describes one type of distortion that results.

Figure 1 above left shows the result of overdriving an amplifier. With low input levels, the amplifier's output signal gives a true reproduction of the input signal. However, when the input signal reaches a certain level, compression begins. Note the point where the output signal is compressed by 1dB, which means that the output level increases 1dB less than the input level. For example, when a 3dB increase in the input level produces a 2dB increase in the output level, as illustrated in Figure 1, 1dB compression has occurred. Specifications for many ampli-

ers and other devices include the 1dB compression point.

When compression begins, IMD results. The IMD begins at a low level and rises rapidly as the input level increases beyond the point of compression.

Test setup

Figure 2 above right shows a simplified setup for measuring IMD produced in the spectrum analyzer. Essentially a sophisticated superheterodyne receiver, a spectrum analyzer can have the same problems as any other superheterodyne receiver. In this test setup, generators A and B generate two continuous wave (CW) tones at RF frequencies. A special combiner (not a simple "T" connector) combines their output and connects it to the spectrum analyzer input. When the input levels to the spectrum analyzer begin to overdrive or overload the analyzer, IMD occurs. Third-

order IMD signals at frequencies of $2A - B$ and $2B - A$ form. ($2A - B$ is two times the frequency of tone A minus the frequency of tone B.)

Highly accurate tests require a more complex test setup than depicted in Figure 2. For example, a low-pass filter should be inserted between each generator and the combiner to suppress harmonics, and a high degree of isolation must be provided to prevent interaction between the two generators. The combiner provides some

(continued on page 68)

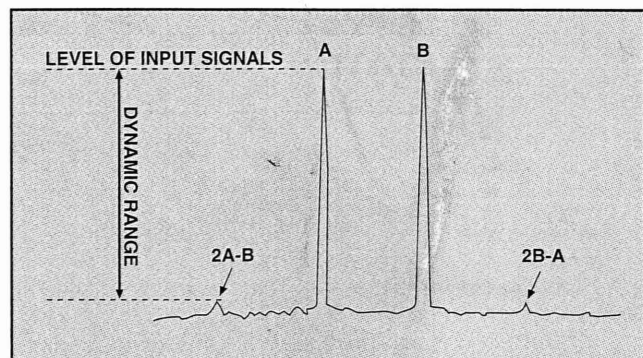


Figure 3. A spectrum analyzer's dynamic range equals the difference of the reference tones (A or B) and the level of the IM signals as the IM signals rise just above the spectrum analyzer's noise floor.

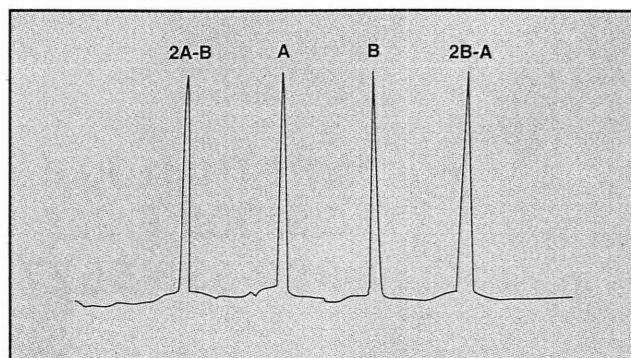


Figure 4. At the theoretical third-order intercept point, the level of the IM signals equals the level of the reference tones. Severe amplifier compression would prevent this condition from occurring in a practical amplifier.

Technically speaking

(continued from page 8)

isolation, and more isolation could be obtained by inserting isolation amplifiers between the generator outputs and the combiner.

Of course, proper padding must be used to maintain a good impedance match throughout. Just remember, the setup shown in Figure 2 is a simplified version of the elaborate setup used in laboratories.

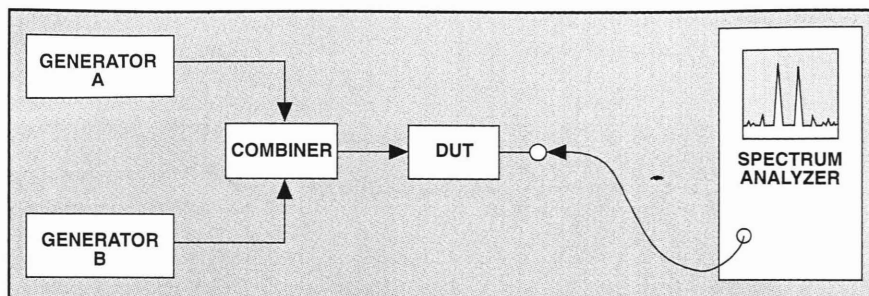


Figure 5. A device under test (DUT) is tested for third-order intermod using this simplified version of the laboratory setup. See text for information about the integrity of the test setup.

The whole purpose is to make sure that any IMD caused by the test setup is suppressed far below the IMD generated in the device under test (DUT)—in this case, the spectrum analyzer.

Dynamic range

Gradually increase the generators' outputs at equal levels until the third-order intermod products just begin to appear above the noise level on the spectrum analyzer display. At this point, the difference between the level of tone A (or B) and one of the intermod products ($2A - B$ or $2B - A$) is called the *dynamic range* of the spectrum analyzer. For example, when the difference between tone A and the $2A - B$ product is 70dB, the dynamic range of the spectrum analyzer is 70dB. (See Figure 3 on page 8.)

Third-order intercept

Once the intermod signals form, the levels of the IM signals increase rapidly as the level of the two tones increases. In fact, the level of the IM signals increases much more rapidly than the level of the fundamental tones (A or B). For this reason, theoretically, a point would be reached where the level of the IM signals would become equal to the level of the fundamental tones, A or B. (See Figure 4 on page 8.) This point is called the *third-order intercept point*. This point occurs when the input tones are increased about 15dB above the point of 1dB compression.

Remember, the third-order intercept point exists only in theory. In practice, an amplifier's compression prevents IMD signals from reaching this point. Regardless, the third-order intercept point is an important specification for amplifiers and receivers. It tells a lot about how the amplifier or device should be operated. Although higher-order intercept points could be specified, third-order intermod is the most prominent and therefore the most troublesome.

Figure 5 above shows how a typical

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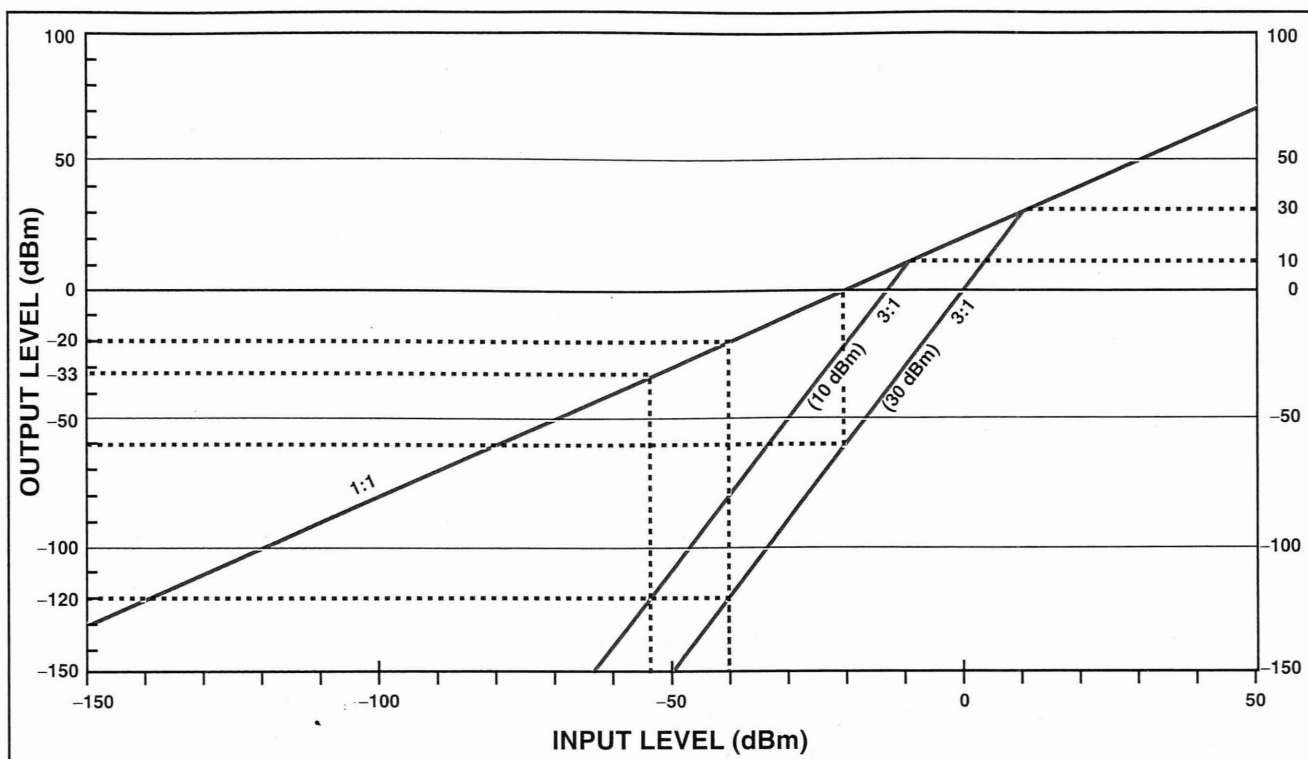


Figure 6. This graph of the third-order intercept point of two amplifiers is useful in determining the amplifier's operating conditions, input/output levels and IM levels. See text for details.

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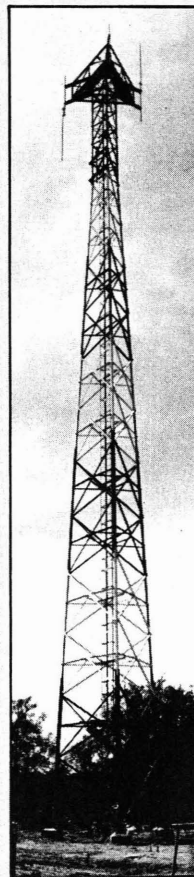
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Technically speaking

DUT is tested. This configuration could represent receiver testing. The spectrum analyzer could be connected to the first IF amplifier's output or another such test point in the receiver. The test would include all circuits between the receiver input and the point of measurement. However, the first active device will be the main determining factor.

Third-order intercept graph

Figure 6 on page 69 shows the third-order intercept points for two amplifiers. Both amplifiers have 20dB gain. One has a third-order intercept point of +30dBm; the other has a third-order intercept point of +10dBm.

Notice that there are three graphs. One represents the amplifier's linear output vs. input—the 1:1 ratio line that means that the output (vertical scale) changes 1dB for each 1dB change in input (horizontal scale). The other two lines representing the intermod signals have a 3:1 slope, which means that, with a 1dB change in the input level, the output level will change by 3dB. The point where the 3:1 graph intersects the 1:1 graph is the third-order intercept point. Notice that one point is at +30dBm, and the other is at +10dBm.

Third-order intercept formulas

These formulas take into account the DUT's gain and are based on the input signal levels.

$$I_P = \left(\frac{3(P_{Ai} + G) - P_3}{2} \right) \quad (1)$$

$$P_3 = 3(P_{Ai} + G) - 2I_P \quad (2)$$

$$P_{Ai} = \left(\frac{(P_3 - 3G - 2I_P)}{3} \right) \quad (3)$$

where

P_3 = power (dBm) in third-order intermod signal.

P_{Ai} = power (dBm) of either tone A or tone B at the input.

G = gain (dB) of amplifier.

I_P = third-order intercept point (dBm).

These formulas represent the levels of tone A and tone B at the point where the spectrum analyzer is connected.

$$I_P = \left(\frac{3P_A - P_3}{2} \right) \quad (4)$$

$$P_3 = 3P_A - 2I_P \quad (5)$$

$$P_A = \left(\frac{(P_3 + 2I_P)}{3} \right) \quad (6)$$

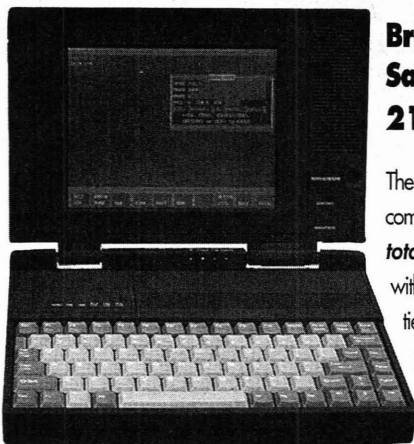
where

P_3 = power (dBm) of third-order intermod signal.

P_{Ai} = power (dBm) of either tone A or tone B.

I_P = third-order intercept point (dBm).

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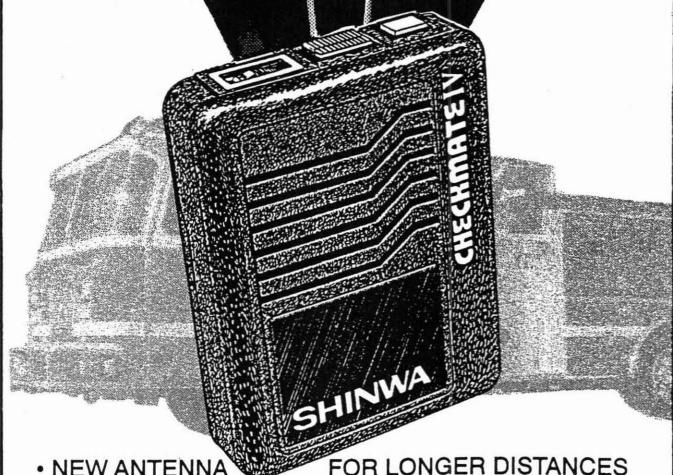


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The graph takes into account the amplifier's 20dB gain. For example, an input level of -40dBm (horizontal scale) corresponds to an output level of -20dBm (vertical scale).

Suppose a receiver has a sensitivity of -110dBm or 0.7μV. To prevent the third-order IM signal from degrading the reception of a -110dBm signal, the IM signal should be suppressed at least 10dB below the desired signal level. In this case, the third-order IM signal should be no greater than -120dBm to prevent any interference to the desired signal.

To determine the maximum input/out-

The IMD ratio can be defined as the difference in the level of the desired tone (A or B) and the level of the third-order IM signal measured in decibels.

put level that the have two tones can reach before causing distortion, a line is drawn from the -120dBm point on the vertical scale to intersect the IM slopes. Then, vertical lines are drawn up to the 1:1 slope and down to the horizontal scale. These lines are shown as dashed lines in Figure 6. The dashed lines intersect the 1:1 slope at the -20dBm point for the +30dBm slope and at -33dBm for the +10dBm slope.

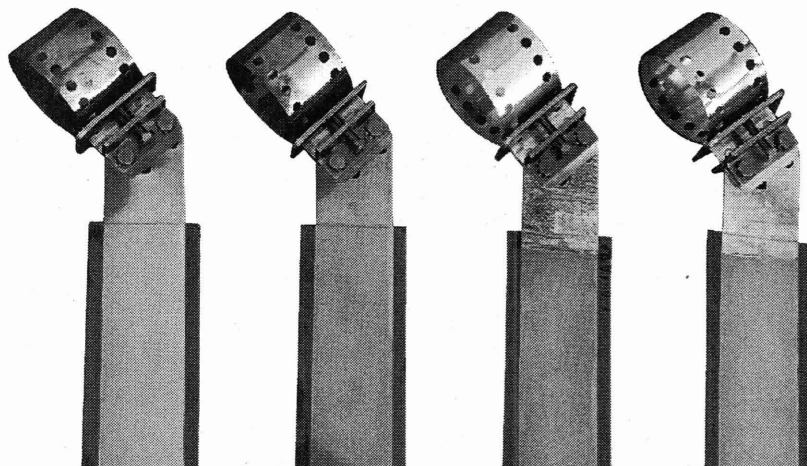
Similarly, the dashed lines intersect the horizontal scale at -40dBm for the +30dBm slope and -53dBm for the +10dBm slope. Thus, for the +30dBm slope, the output level of tone A or B should be no greater than -20dBm, and the input level to the amplifier should be no greater than -40dBm.

For the +10dBm slope, the output level of tone A or B should be no greater than -33dBm, and the input level should be no greater than -53dBm. If the input/output levels exceed these limits, the third-order IM signal will exceed the -120dBm level and will degrade the -110dBm desired signal.

IMD ratio

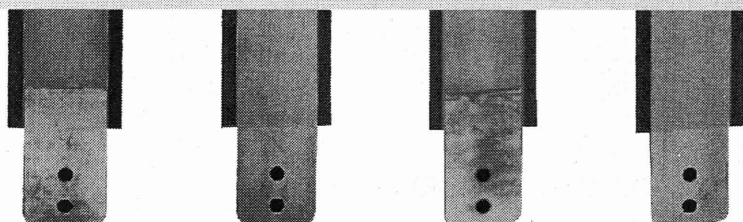
The IMD ratio can be defined as the difference in the level of the desired tone (A or B) and the level of the third-order IM signal measured in decibels.

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Technically speaking

The IMD ratio must be specified at a given input or output level because the IMD ratio becomes worse as the input/output level increases.

Take the case of the 20dB amplifier with the +30dBm third-order intercept point. When the output level of tone A or B is -20dBm, the level of the third-order IM signal is -120dBm. This difference represents an IMD ratio of $(-20) - (-120) = 100\text{dB}$.

Now, look at the case for an output level of 0dBm. As shown by the dashed line, the third-order IM signal is -60dBm. Thus, the IMD ratio is now 60dB. The ratio gets worse fast as the input/output level increases!

Formulas

You can use formulas listed on page 70 to make calculations involved with third-order intercept. Formulas 1-3 take into account the DUT's gain and are based on the *input* signal levels. This means that the spectrum analyzer used to study the intermod is connected to the *output* of the device, and the signal levels used in the formulae (P_{Ai}) are the levels of the tones (A or B) at the input to the device.

Formulas 4-6 represent the levels of the tones (A or B) at the point where the spectrum analyzer is connected. These examples demonstrate the practical use of the formulas.

Example: The DUT is an RF amplifier with 20dB gain and a +20dBm third-order intercept point. The level of the third-order IM signals must be kept to no more than -60dBm. What is the maximum level of tone A or B short of causing IM signals greater than -60dBm?

Solution: Use Formula 3 on page 70 and substitute values from the example.

$$P_{Ai} = \left[\frac{-60 - 3(20) + 2(20)}{3} \right]$$

$$= \left(\frac{-80}{3} \right)$$

$$= -26.7\text{dBm}$$

Example: The DUT is an RF amplifier. A spectrum analyzer connected to the output shows the third-order IM signal at a level of -60dBm and the level of tone A at -10dBm. What is the third-order intercept point?

Solution: Use Formula 4 on page 70 and substitute values from the example.

$$I_P = \left[\frac{3(-10) - (-60)}{2} \right]$$

$$= \left(\frac{+30}{2} \right)$$

$$= +15\text{dBm}$$

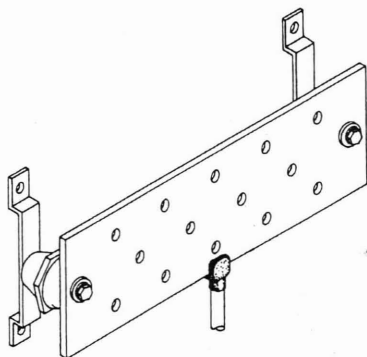
Next time, we will look at intermodulation from a different perspective and learn how to take advantage of the behavior of the third- and higher-order intermodulation process to gain an advantage in suppressing it. *Stay tuned!*

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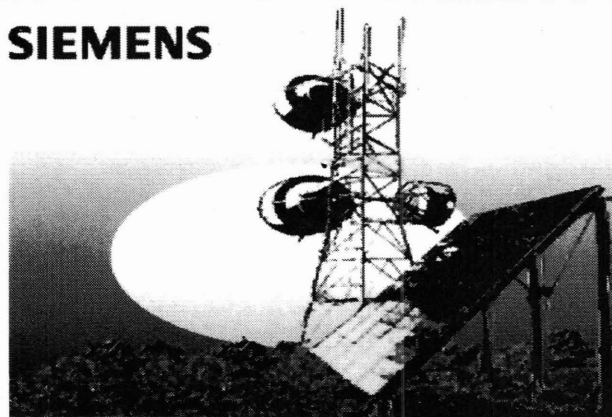
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More about intermodulation

By Harold Kinley, C.E.T.

Last month, we discussed third-order intercept and two-tone intermodulation testing using two equal-level, single-tone signals (A and B) to produce the intermod (IM) signal. Though two equal level tones (A and B) are used in third-order intermod testing of amplifiers and receivers, in the real world the mixing signals are rarely of equal level. In fact, one of the mixing signals usually has a much higher level than the other(s).

This discussion is limited to two-signal, third-order intermodulation. Though higher orders of intermod are often encountered, and more than two mixing signals can be involved, the third-order, two-signal IM is usually the most troublesome.

Calculating IM level

The formulas in the box above can be used to determine how unequal-level tones (A and B) produce unequal IM signals, $2A - B$ and $2B - A$. (The formulas also work for equal-level, two-tone signals). Furthermore, the power in the two IM signals can be determined from the power level of the two-tone mixing signals, A and B, and the third-order intercept point of the amplifier or other device under test (DUT).

Kinley, a certified electronics technician, is regional communications manager, South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985. He is a member of the Radio Club of America.

Formulas for calculating IM levels

$$P_{(2A-B)} = 2P_A + P_B - 2P_{I3}$$

$$P_{(2B-A)} = 2P_B + P_A - 2P_{I3}$$

where: $P_{(2A-B)}$ = power level of IM signal in dBm

$P_{(2B-A)}$ = power level of IM signal in dBm

P_A = power level of signal A in dBm

P_B = power level of signal B in dBm

P_{I3} = power level of third-order intercept in dBm

Figures 1 and 2 below show how two unequal tones, A and B, produce two unequal IM signals, $2A - B$ and $2B - A$, at the output of an amplifier with a third-order intercept point of +30dBm. In Figure 1, tone A is at +10dBm, and tone B is at -10dBm. The $2A - B$ IM signal is at -50dBm, whereas the $2B - A$ IM signal is at -70dBm. Figure 2 shows a reverse of the levels of the two tones: A is at -10dBm, and B is at +10dBm. Notice how the two IM signals are reversed from Figure 1. From this comparison, we can see that the higher of the two IM signals, $2A - B$ or $2B - A$, is always nearest the higher level tone, A or B.

The graph

Figure 3 on page 79 is handy for calculating third-order IM levels when the levels of the two tones, A and B, are known. The graph is for an amplifier or other device with a third-order intercept point of +30dBm; however, it can be used for amplifiers with third-order intercept points other than +30dBm by applying an adjustment factor.

There are two horizontal scales at the top and bottom of the graph. The top horizontal scales are where the IM signal level is read. The bottom horizontal scales represent the level of the second-order term, A or B. For example, if the IM is formed from $2A - B$, then the term A is the second-order term, and the term B is the first-order term, and the resulting IM level found on the top horizontal scale would represent the $2A - B$ IM signal level. Conversely, if the IM is formed from $2B - A$, then the B term is the second-order term, and the A term is the first-order term, with the resulting IM level found at the top horizontal scale representing the $2B - A$ IM level.

The right vertical scale of the graph represents the first-order term of the IM signal. The major divisions represent 10dB increments.

The left vertical scale of the graph can be used to adjust for third-order intercept points other than +30dBm. The

(continued on page 79)

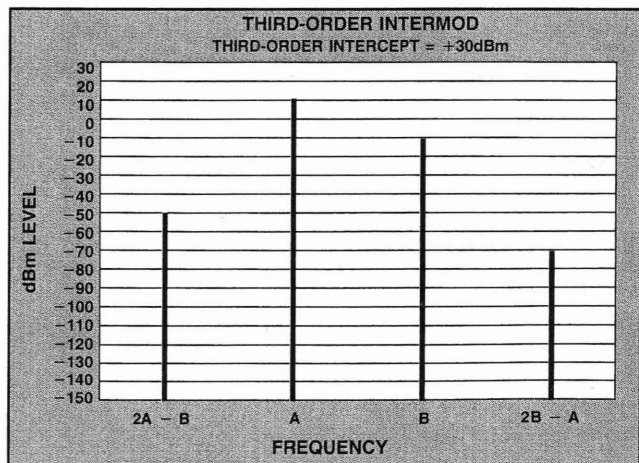


Figure 1. Signals A and B mix to form IM signals $2A - B$ and $2B - A$. Note that the higher-amplitude IM signal is nearest to the higher-amplitude mixing signal, A.

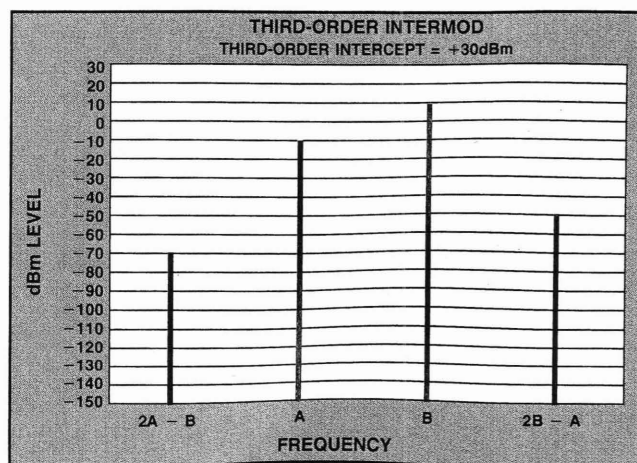


Figure 2. The mixing signals, A and B, are reversed from Figure 1. Now, the higher-amplitude mixing signal is B. Again, the higher-amplitude IM signal ($2B - A$) is nearest to the higher amplitude mixing signal, B.

Technically speaking

(continued from page 8)

major divisions represent 5dB increments. For the +30dBm third-order intercept point, no adjustment is necessary.

Using the graph

A few examples illustrate how to use the graph. Looking back at Figure 2, let's use the graph to determine the levels of the $2B - A$ and the $2A - B$ IM signals. In the case of $2B - A$, B (+10dBm) is the *second-order* term, so locate B on the lower horizontal scale. Now move up as shown by the red dashed line to intersect the graph at point 1. Now, to take into account the *first-order* term, A at -10dBm, move down (down for negative, up for positive) 10dB to point 2 and back over to intersect the graph at point 3. From point 3 move up to the horizontal scale to find the level of the $2B - A$ IM signal is -50dBm.

The level of the $2A - B$ IM signal is found similarly. This is shown by the blue dashed line on the graph. This time, the *second-order* term is A , and the *first-order* term is B . The *second-order* A term (-10dBm) is located on the bottom horizontal scale. From here, move up to intersect the graph at point 1. Now, to take into

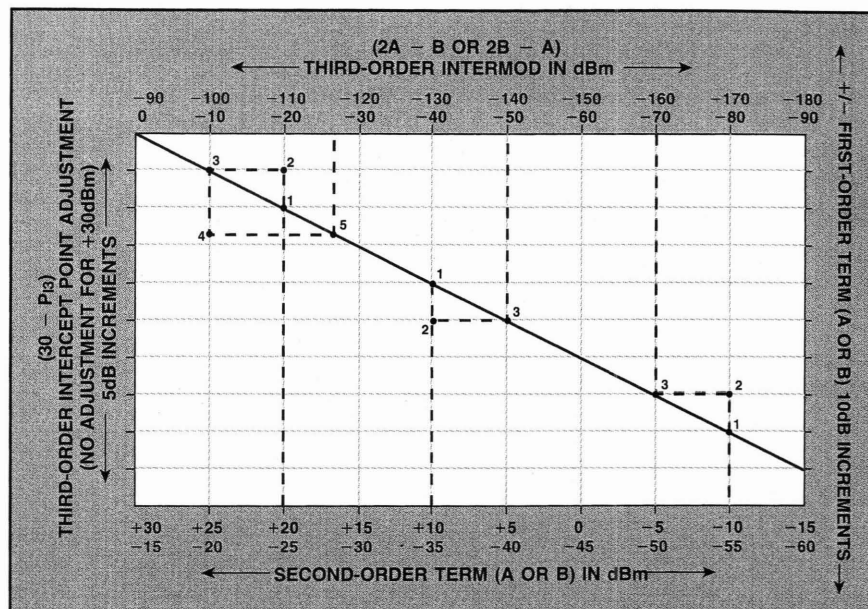


Figure 3. This graph can be used to determine the amplitude of third-order, two-signal IM signals such as $2A - B$ or $2B - A$. See text for instructions on using the graph.

account the *first-order* B term (+10dBm), move up 10dB to point 2 on the graph. Now move back over to intersect the graph at point 3 and up to the top horizontal scale

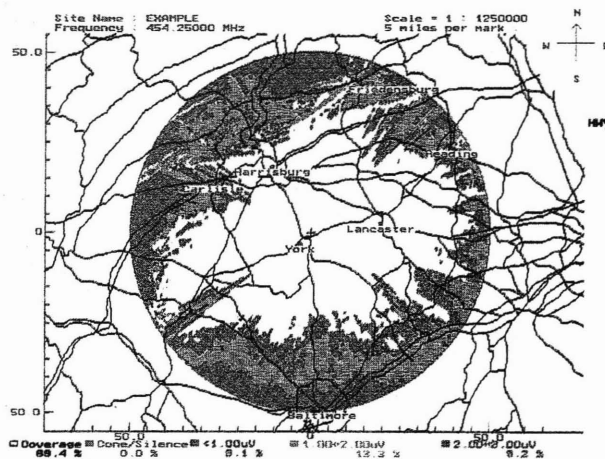
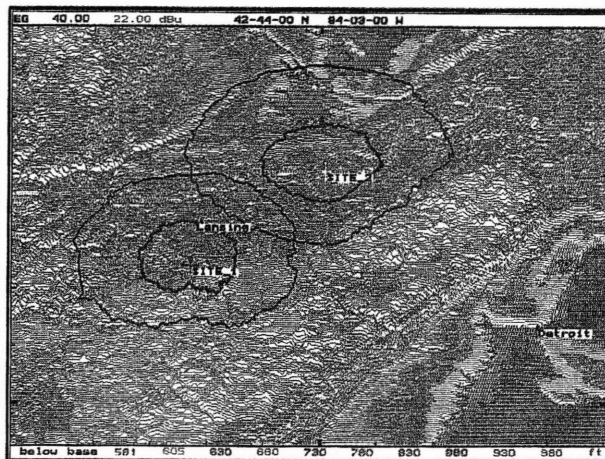
at -70dBm. Thus, -70dBm is the level of the third-order ($2A - B$) IM signal.

Here is an example using an amplifier with a third-order intercept point of



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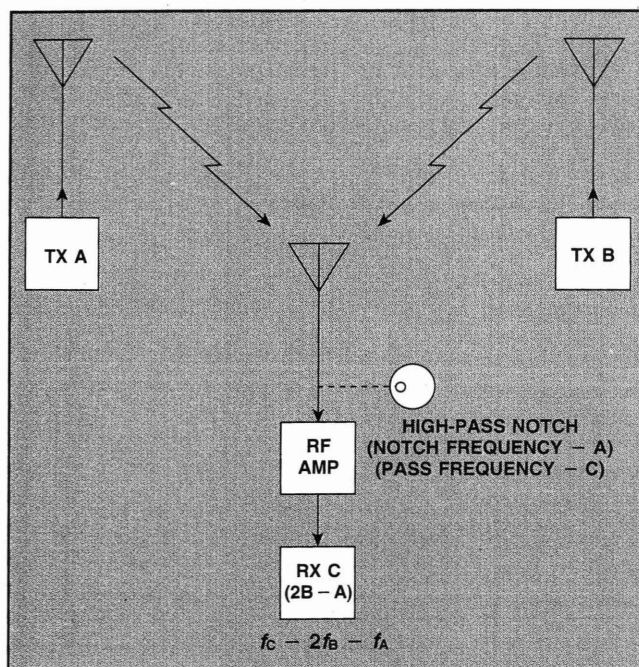


Figure 4. (to the left) Signals from transmitters A and B are mixing in receiver C to produce an IM "hit" on receiver C. ($C = 2B - A$). A high-pass notch filter shown in dashed line can be connected to the line at the receiver to suppress the IM signal to a non-interfering level.

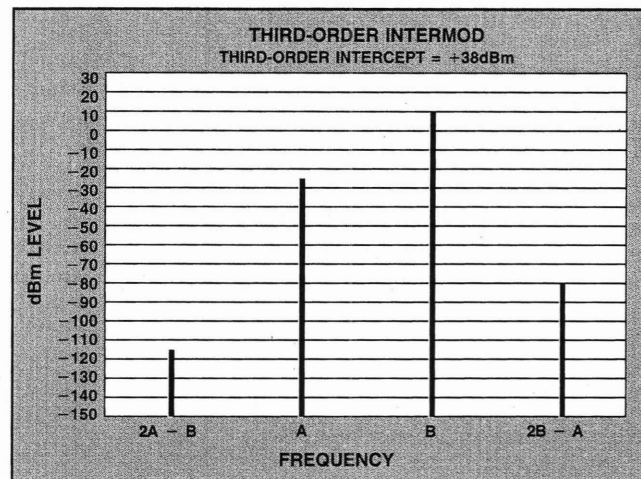


Figure 5. These signals are present at the output of the RF amplifier of Figure 4. Note that the level of the $2B - A$ IM signal is at an amplitude of -80dBm . This would override the desired signal at a level of -106dBm .

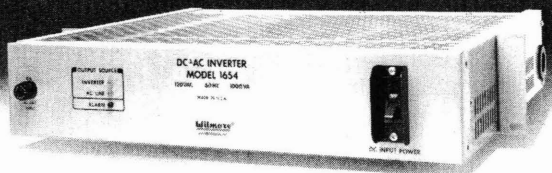
$+38\text{dBm}$ (a high-quality amplifier). The signal levels at the output of the amplifier are as follows: signal A is -25dBm , and signal B is $+10\text{dBm}$. What is the level of

the $2A - B$ IM signal? First, locate -25dBm on the lower horizontal scale, and from there move up to the graph at point 1. Now, to account for the $+10\text{dBm}$ level of signal

B, move upward 10dB to point 2. Now move back to the graph line at point 3. Because the third-order intercept point of this amplifier is $+38\text{dBm}$ instead of

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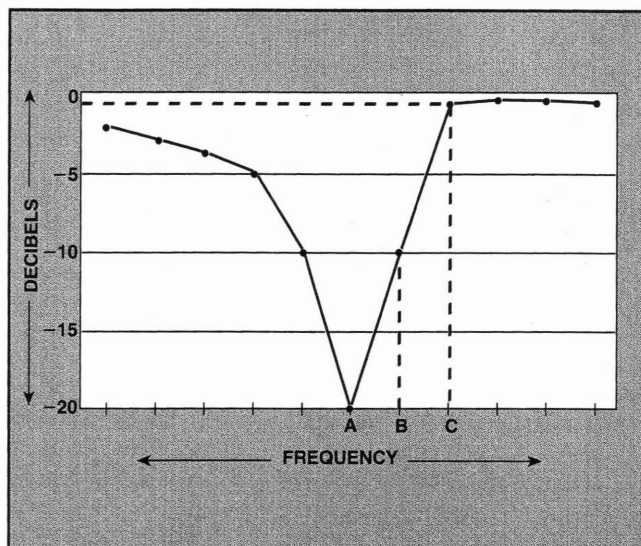


Figure 6. The frequency response of the cavity filter used in Figure 4 is shown here. Note that the signal from transmitter A is reduced by 20dB at the input to the receiver amplifier, whereas the signal from transmitter B is reduced by 10dB. The insertion loss at receiver frequency C is less than 1dB.

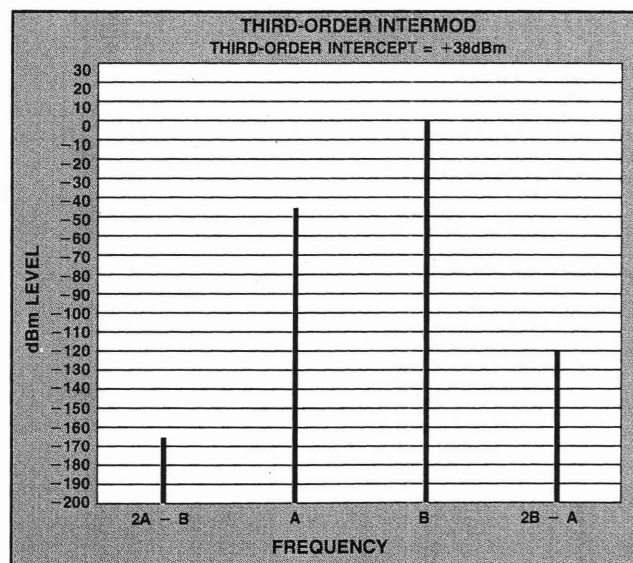


Figure 7. This is the level of the various signals at the Figure-4 RF amplifier's output after the high-pass notch filter is connected. Note that the 2B - A IM signal is now reduced to -120dBm, resulting in little or no interference to the desired signal.

+30dBm, we must adjust for the difference. The adjustment will be $30 - 38 = -8\text{dB}$. So, from point 3 on the graph, move down 8dB according to the left

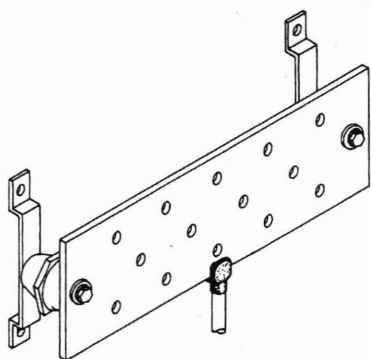
vertical scale to point 4. Next, move over to the graph line to point 5. From here, move vertically to the top horizontal scale to a reading of -116dBm. This is

the level of the 2A - B IM signal.

Receiver IM suppression

Figure 4 on page 80 shows signals from

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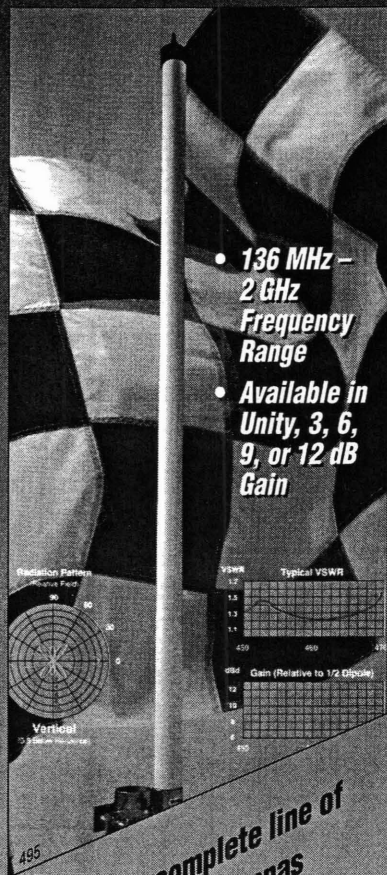
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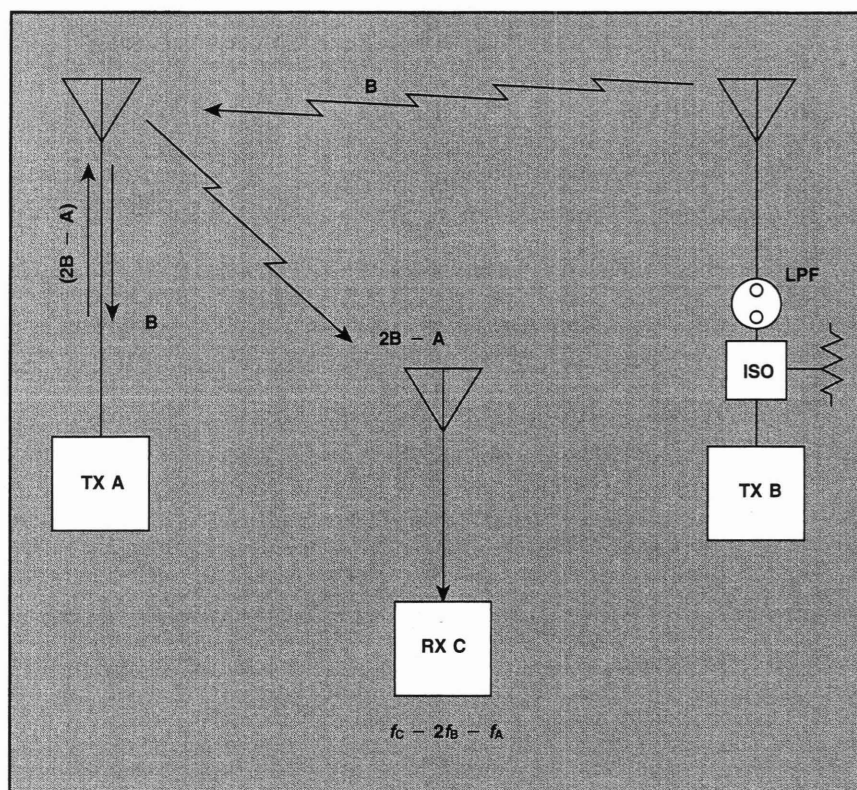


Figure 8. In this situation, the IM signal is being produced in the class C RF power amplifier of transmitter A. The IM signal can be suppressed by installing a proper isolator on the transmission line at transmitter A.

transmitter A and B mixing in the front end of receiver C. The frequency combination of $2B - A$ happens to be right on the receiver frequency C. The levels of the various signals as they appear at the output of the RF amplifier of receiver C are shown in Figure 5. Suppose that it is desired that the receiver be able to receive a $0.35\mu\text{V}$ signal on frequency C, $(2B - A)$ with minimum degradation caused by the IM signal. Suppose further that the RF amplifier's gain is 10dB.

This means that, with a $0.35\mu\text{V}$ (-116dBm) signal at the input of the receiver RF amplifier, the output of the desired signal from the amplifier will be -106dBm . Note that the level of the $2B - A$ IM signal at the RF amplifier output is -80dBm . This is 26dB above the level of the desired signal. A high-pass notch filter is placed on the line at the receiver RF amplifier input to suppress the IM signal to a non-interfering level. (See dashed line in Figure 4.)

The frequency response of the high-pass notch filter is shown in Figure 6 on page 81. Note that the level of signal A is attenuated by 20dB, and the level of signal B is attenuated by 10dB. So what happens to the level of the IM signal at the output of the RF amplifier?

Figure 7 on page 81 shows the levels of

the various signals at the RF amplifier output after a high-pass notch filter is installed on the line before the receiver input. Note that the level of the interfering IM signal is now -120dBm . Remember, the level of the desired signal was -106dBm . So, even with the slight insertion loss of the filter at the desired frequency, the IM signal is now suppressed approximately 13dB below the desired signal and should not cause any significant degradation of the desired signal.

Transmitter IM suppression

Figure 8 above shows a situation in which transmitter-produced intermodulation is producing an IM hit on an on-site receiver.

Note that an isolator and low-pass filter are inserted in the line between transmitter B and the antenna. The isolator provides a high degree of suppression for any signal traveling from the antenna back toward the transmitter. The signal traveling in this direction is "dumped" into the load resistor and dissipated as heat. The low-pass filter serves to suppress the harmonics that might be generated by the isolator.

Note that transmitter A has no isolator in the line. The signal from transmitter B at the antenna can come down into the transmitter's class C power amplifier and

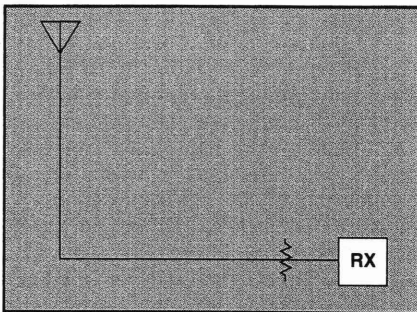


Figure 9. A simple attenuator pad can be inserted in the line at a receiver to determine whether the IM signal is receiver produced or radiated from another source. See text for details.

mix with the signal of transmitter A to produce an IM signal that will interfere with receiver C. To suppress the IM signal to a non-interfering level, an isolator should be installed in the line of transmitter A along with a low-pass filter to suppress any harmonics generated by the isolator.

Receiver-produced IM

Before setting out to suppress IM interference, you must first know where it is being produced. Is it being produced in the front end of the receiver, or is it being produced externally and radiated to the receiver antenna from another source? One simple way to decide is to use an attenuator pad connected before the receiver input. (See Figure 9 above.) If the IM signal is third-order and is being produced inside the receiver, the attenuation of the IM signal will be three times the pad attenuation value.

For example, if a 3dB pad is inserted into the line, the receiver-produced third-order, two-signal IM will be reduced by 9dB. Unfortunately, the desired signal will be attenuated by 3dB, but we gain the advantage of a 6dB difference with the IM signal. A fifth-order, two-signal IM product will be reduced by 15dB. If the IM signal drops less than the expected amount but more than the pad attenuation, the IM signal is probably a combination of receiver-produced and radiated IM.

There is much more to be written about intermod than can be included in this space. This column has only scratched the surface; however, there is a book dedicated to the subject of intermodulation control. The book, *Intermod Control*, was written by William F. Lieske Sr., president of EMR, Phoenix, AZ. Lieske is a recognized authority on the subject of intermodulation control. The book is available from Wiesner Publishing. Individual copies can be ordered for \$10 postpaid with a check or money order made out to Wiesner Publishing, Attention Mike Peterman, 7009 S. Potomac, Englewood, CO 80112;

tel. 303-397-7600; fax 303-397-7619.

A large, functional version of the graph in figure 3 is available from the author for \$2.00 post-paid.

A computer program that is useful in calculating IM hits and third-order, two-signal IM is available from the author for \$12.50 plus \$2.50 for shipping and handling. These programs run from MS-DOS on a PC and are available on a 3.5" or 5.25" floppy diskette. You may write the

author at 204 Tanglewylde Drive, Spartanburg, SC 29301-2949.

Stay tuned!

References

- (1) Christiansen, Donald, and Donald G. Fink, *Electronics Engineer's Handbook, 3rd edition*, 1989, McGraw-Hill, New York
- (2) Lieske, William F. Sr., *Intermod Control*, 1984, Wiesner Book Company, Littleton, CO.



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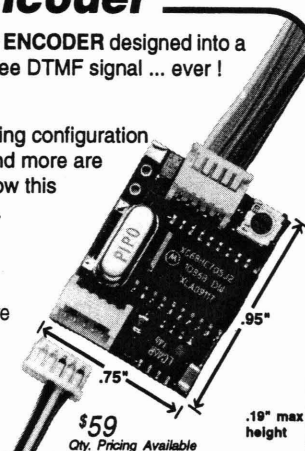
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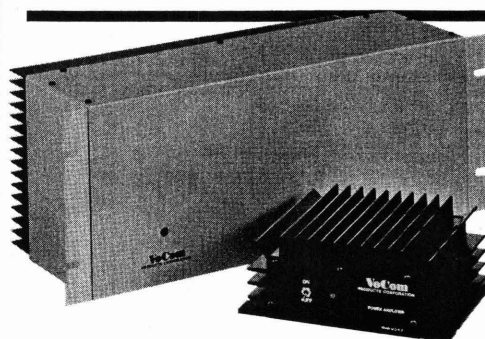
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How F-TDMA can improve private land mobile radio

Frequency-and-time-division multiple-access offers a path that allows refarming of existing spectrum below 470MHz to 12.5kHz spacing by on-center channel migration while delivering a 100% increase in efficiency.

By John Yoon
and Barbara Baffer

The digital revolution surrounds us. Compact discs have overtaken audio cassettes and made vinyl records candidates for antique shops. Television manufacturers have invested six years, since 1987, testing alternative

architectures.

Ten years ago, the early personal computers were gaining acceptance. Now, multimedia personal computers integrate video, text and sound for learning and research as well as games.

Digital cellular telephone systems operate in Europe, Japan, the United States and Canada.

Land mobile radio is on the verge of

exploiting this same digital technology to deliver increased capacity, improved performance and enhanced functionality. A sound foundation is needed before one can build a good quality house. The same is true for fun-

Yoon is manager of public safety marketing, and Baffer is a regulatory analyst at Ericsson GE Mobile Communications, Lynchburg, VA.

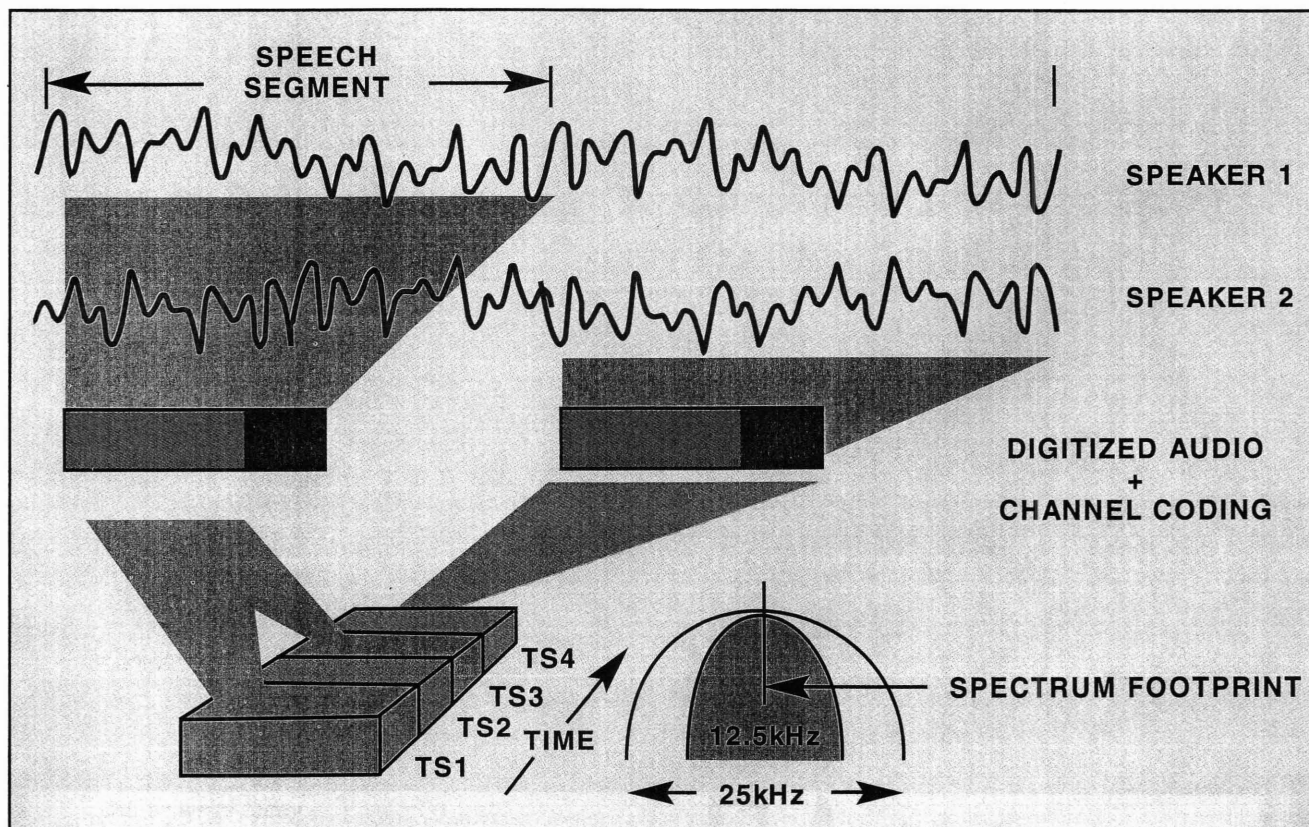


Figure 1. Time-division multiple-access (TDMA) technology allows communications among multiple users by allocating unique time slots to different users. This technique increases capacity by time-multiplexing users on the same RF channel. The shadings represent two distinct user groups. TDMA leaves the channel intact and uses a time reference to keep the data organized within the channel.

damental technology decisions necessary to build an advanced communication system. The architecture is critical to the delivery of promised advantages to land mobile users.

For example, in the mid '80s, the Japanese rushed to adopt a high-definition television (HDTV) protocol. They selected an analog standard because of the speed of product development. They were the first to offer improved video images.

Unfortunately, no other country has adopted that standard, so the Japanese have been unable to capitalize on being first. Other countries, such as the United States, elected to invest in a digital standard that provides not only improved video and sound, but opportunities for television monitors to serve as the interface to interactive services and to advanced home energy control devices, for example.

Meanwhile, the Japanese will have to trade out their broadcasting and receiver equipment again to enjoy these benefits.

This article explains the radio communications benefits promised with digital technology. Time-division multiple-access (TDMA) technology, a spectrum-efficient implementation for digital land mobile radio, is explained. The characteristics of TDMA are contrasted with another approach, fre-

Digital technology improves voice and data communications performance . . .

quency-division multiple-access (FDMA). A proposal for TDMA on narrowband, 12.5kHz channels, termed F-TDMA, is described along with its associated advantages.

Digital advantages

Digital technology offers increased

spectrum efficiencies and advanced features.

Specifically, digital technology offers the following:

► *Improved performance*—Digital technology improves voice and data communications performance, thanks to sophisticated digital signal processing (DSP) units in mobile and portable equipment.

► *Security*—Digitized speech provides inherent protection and a significant level of privacy and security from conventional scanners.

Sophisticated encryption technology can enhance this security for sensitive tactical communications.

► *Advanced features*—Digital technology allows the design of a signaling structure with the flexibility to handle several new features, such as:

- data-over-voice signaling.
- automatic power control of subscriber units.
- high-throughput data and dynamic bandwidth applications.
- advanced networking and net-

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F-TDMA questions and answers

Ericsson-GE responds to some questions about F-TDMA:

MRT: How is the digital technology developed as F-TDMA for private land mobile radio (PLMR) the same as the digital technology used in cellular mobile telephone systems?

EGE: By "the same digital technology," we mean the digital technology used in all kinds of products.

With respect to using digital cellular telephone technology for PLMR, advancements in very large scale integrated circuit (VLSIC) technology have provided an avenue for digital signal processors (DSPs) to be developed. DSPs can be programmed to modify or add functions within a system without redesigning a custom IC. In this sense, digital technology can provide PLMR communications with new and powerful capabilities.

The way we use the term F-TDMA, the F defines a channel bandwidth of 12.5kHz vs. the current 25kHz bandwidth. The key in this design lies in adapting the digital cellular TDMA for PLMR use.

Fortunately, the use of digital components allows most of the changes to be accomplished via software; therefore, it makes sense to say that F-TDMA is exploiting the same technology as digital cellular. This does not mean that F-TDMA has the same system architecture as digital cellular, though. Technology should not be confused with system architecture.

Although a 12.5kHz, 2-slot F-TDMA system is not yet in production, a prototype was demonstrated at the Associated Public-Safety Communications Officers (APCO) national conference in August in New Orleans. A similar 12.5kHz, 2-slot F-TDMA concept is being defined for the Trans-European Trunked Radio (TETRA) RES-6 standard.

MRT: You have listed a number of advantages that digital technology has to offer PLMR, such as improved performance, security, advanced features and spectrum efficiency. Are these advantages specific to F-TDMA, or can they be delivered by other digital architectures?

EGE: Digital technology ad-

vances most, if not all, radio architectures.

MRT: Will scanner manufacturers eventually make receivers that can be used to listen to digital communications? If so, how can digital PLMR users protect their communications from competitors and criminals?

EGE: It is fairly easy to monitor analog FM communications with a scanner. Digital communications cannot be monitored with the same type of scanner.

Digital scanners, if they ever become available, are likely to carry much higher prices because of the more complex circuitry required compared to analog scanners and because of the required economies of scale.

First, casual listeners and scanner hobbyists who make no use of information in conversations they overhear pose no threat.

Second, the average competitor or criminal who eavesdrops only when it is easy and inexpensive, as with an analog scanner, will not be able to listen to digital communications.

Third, if digital scanners become available, digital radio communications technology offers a low-cost software feature for a low level of privacy and security sufficient to prevent monitoring by stock scanners.

Fourth, sensitive business communications that might be the target of industrial espionage and tactical public safety communications that might be the target of sophisticated criminals can be protected with digital encryption standard (DES) encryption.

DES encryption requires such enormous resources and so much time to decode without first knowing the key that the information it protects usually is without value by the time it is decoded.

MRT: One of the advantages of digital communications you describe is dynamic bandwidth. Does that mean that the bandwidth changes from time to time?

EGE: Yes, although the bandwidth would not change during a single call.

Dynamic bandwidth might be referred to as *bandwidth on de-*

mand. In an F-TDMA system, the bandwidth never would be greater than what current regulations specify.

Here are two examples of how dynamic bandwidth applies to an F-TDMA system:

(1) An F-TDMA system with dynamic bandwidth equipment offers a migration path from the current "maximum" bandwidth (25kHz) to a "one-half maximum" bandwidth (12.5kHz).

Imagine a single base station that can talk to old (maximum bandwidth) and new (one-half maximum bandwidth) radios. With such a base station, existing radio units do not have to be replaced simultaneously with new units.

(2) Using F-TDMA on a 25kHz channel allows four users to conduct separate communications simultaneously. This type of system configuration is permitted under current FCC rules.

MRT: Doesn't the promise of digital technology transfer lie in its ability to deliver a high information transfer rate without increasing bandwidth?

EGE: This is correct; nevertheless, some data applications such as transmitting fingerprint images require a wider bandwidth.

F-TDMA can satisfy requirements for wider bandwidths by providing the flexibility to combine bandwidths of multiple time slots. As technology evolves, the amount of bandwidth required will decrease. Digital technology will promote this bandwidth reduction further, along with the ability to deliver a higher information transfer rate.

Despite the current need for wider bandwidths for specific applications, digital technology not only provides higher information throughputs, but also uses vocoder algorithms to compress voice (which is spectrally inefficient) into a smaller bandwidth.

MRT: Why aren't high channel data rates possible, such as 16kb/s in a 12.5kHz channel? Why can't a 12.5kHz FDMA channel also operate at a 16kb/s data rate?

EGE: Data rates are a function of modulation techniques and occupied bandwidth (mask of signal).

The 16kb/s data rate fits within

the current FCC Part 88 mask using a 4-phase modulation scheme. A 4-phase modulation scheme ($\pi/4$ DQPSK) is used because this demodulation technique can be implemented easily using today's DSP technology.

More complex modulation schemes, such as 8 QAM and 16 QAM, typically require linear amplifiers to maintain energy within the mask and signal characteristics. Linear amplifier technology is advancing at a rate that should allow these more complex modulation schemes to be applied to digital PLMR in two years or so. Where 4-phase modulation can achieve a 16kb/s data rate, 16 QAM (phase) modulation can increase the rate to approximately 32kb/s.

Another way to improve channel capacity without increasing bits/sec/Hz is by further compressing voice signals into a vocoder. Current vocoder technology compresses voice approximately 4kb/s, but it requires approximately 3kb/s of error correction.

Although this represents significant progress from where technology was a few years ago, improvements in vocoder technology are forthcoming. Higher data rates are possible, but modulation scheme selection is limited by linear amplifier technology and demodulation DSP complexity.

MRT: Isn't the real issue information transfer rates vs. bandwidth with frequency reuse? Information about adjacent channel protection ratios and co-channel reuse or desired-to-undesired signal ratios would be useful.

EGE: Yes, frequency reuse is a factor in estimating spectral efficiency.

Our F-TDMA approach uses the same frequency reuse plan that exists today. No new frequency coordination plans are required.

F-TDMA produces 100% spectral efficiency improvement over 12.5kHz FDMA systems and a four times improvement over 25kHz analog systems. The unique feature of F-TDMA is its allowance for transparent migration on existing frequencies without performing a new frequency coordination plan.

work-wide features.

► **Spectrum efficiency**—Digital technology, when implemented properly, promises increased spectrum efficiency, i.e., the ability for higher numbers of users to share the same spectrum (higher capacity) or the ability to pass higher quantities of data (higher *throughput*) through a radio system. With available technology, capacity gains of 100% within 12.5kHz channels and at least 300%-440% in 25kHz channels are possible. Data rates exceeding 30kb/s in 25kHz channels or 16kb/s in 12.5kHz channels are possible.

Spectrum congestion is a fact of daily life to many land mobile radio users. The FCC has initiated a *Notice of Proposed Rulemaking* in PR Docket No. 92-235 to *refarm* frequencies below 512MHz to make additional capacity available. Although it is not required by the refarming initiative, digital technology provides a way to gain spectrum efficiency while im-

proving service quality and functionality. The alternative of new, *green* spectrum is not available to meet the demand for private land mobile radio services.

Competing technologies

TDMA is a relatively new entrant into mobile communications and is challenging traditional approaches.

TDMA historically has been projected for 25kHz or 30kHz land mobile channels. A number of countries, including the United States, have adopted or plan to implement 12.5kHz channels in some frequency bands. This has led our company to develop TDMA for operation on narrowband, 12.5kHz channels, termed F-TDMA.

For cellular telephones, TDMA systems are available and operating successfully in many of the large North American markets. Several of the largest specialized mobile radio (SMR) system operators have announced that they will install TDMA technology to

gain interoperability and to facilitate wide-area roaming.

The authors' company has supported TDMA for private radio since 1989, when the access technology for private radio first came to be publicly debated in the FCC's *Further Notice of Inquiry* in Gen. Docket No. 88-441.

TDMA uses a common channel for communications among multiple users by allocating unique time slots to different users. This technique increases capacity by *time multiplexing* users on the same RF channel.

The shadings in Figure 1 on page 44 represent two distinct user groups. Each group uses the channel resources on a time-shared basis.

In effect, TDMA leaves the channel intact and uses a time reference to keep the data organized within the channel. The users are not aware that they are sharing an RF channel. Cellular telephone, landline telephone, digital microwave and other electronic switches use a similar technique to multiplex conversations.

Frequency-division multiple-access (FDMA) narrows the channels from 25kHz or 30kHz to 12.5kHz. In some cases, adjacent channel interference prevents simply splitting a 25kHz channel into two 12.5kHz channels.

"Multiple access" is a misnomer in the case of FDMA. In fact, unless the user is licensed to adjacent channels, only one user can occupy the channel at a time.

Incorporating the advantages of narrowband channels with the capacity and flexibility of TDMA, F-TDMA provides two-slot TDMA on 12.5kHz channels, as shown in Figure 2 to the left. Capacity is absolutely doubled on each narrowband channel.

Digital modulation

In digital technologies, information is transferred through the air using electromagnetic energy, and the properties of these waves are determined by a modulation scheme.

Modulation techniques often determine the gross bits per hertz. One way digital technologies increase spectrum efficiency is through the modulation scheme. A linear modulation scheme, *differential quadrature phase-shift keying* (DQPSK) is a good match for some

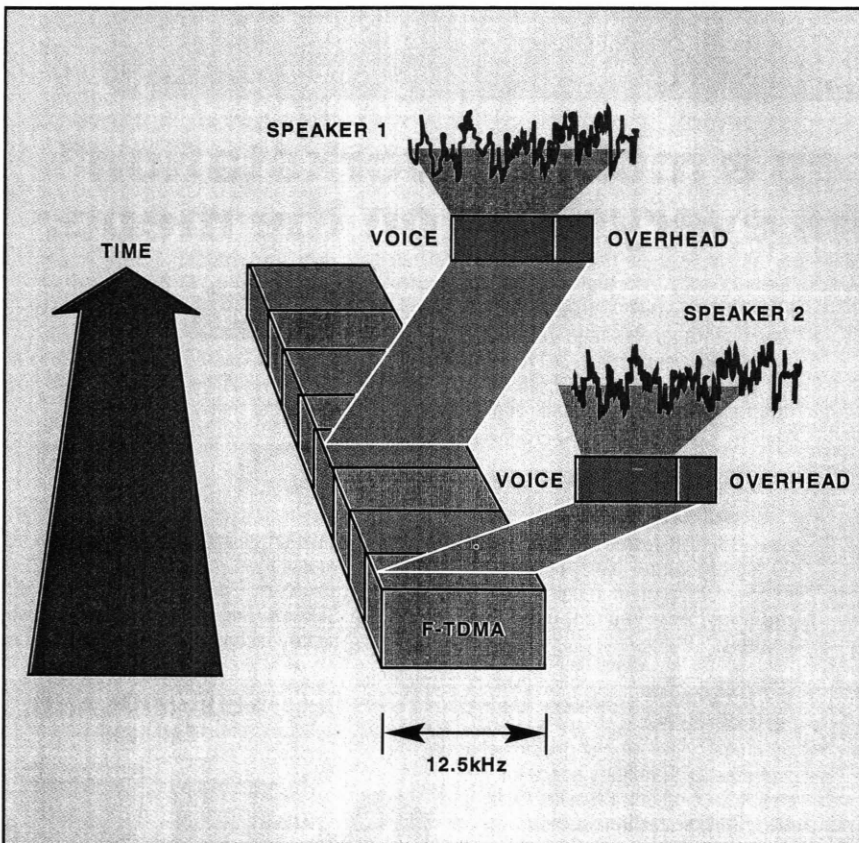


Figure 2. Incorporating the advantages of narrowband channels with the capacity and flexibility of TDMA, F-TDMA provides two-slot TDMA on 12.5kHz channels. Capacity is doubled on each narrowband channel.

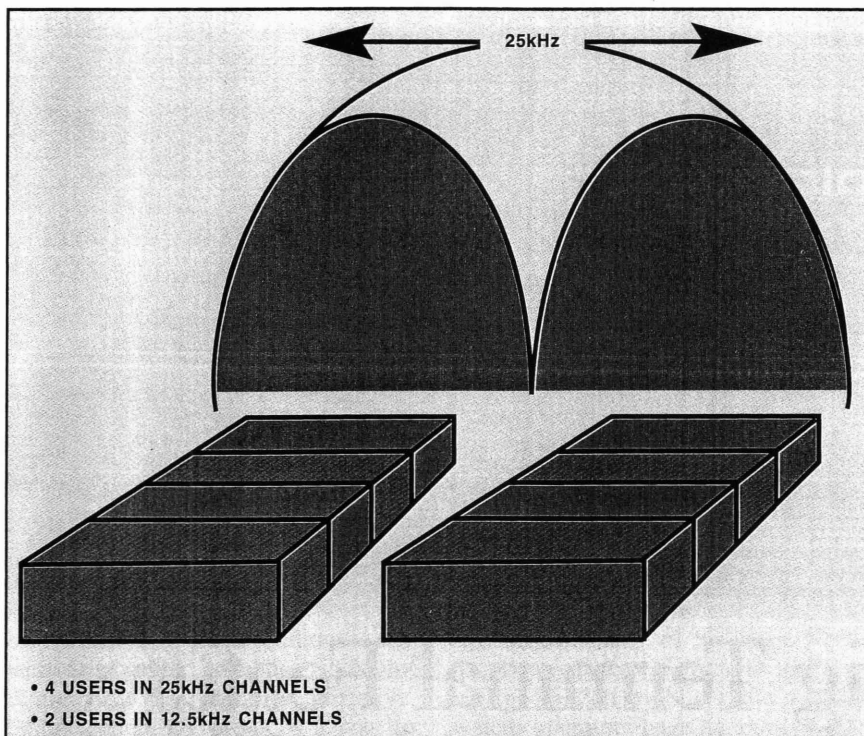


Figure 3. The refarmed spectrum allows two TDMA carriers 12.5kHz apart on a 25kHz channel. This method allows four users on the equivalent of one current channel.

TDMA applications.

DQPSK-modulated F-TDMA can be used in the downlink and can accommodate 16kb/s in a 12.5kHz channel. DQPSK modulation is selected for the downlink communications path to allow simulcast to be implemented. DQPSK represents digital data by shifting the relative phase of the carrier to represent digital symbols. The uplink path, mobile-to-base, will be supported by *continuous phase partial response modulation* (CPM) that eliminates the need for linear or quasi-linear amplifiers in hand-held equipment.

Vocoders

Voice coders (vocoders) are used in digital systems to convert analog speech into digital signals for transmission.

At the receiving end, the vocoder translates the digital signal back into analog speech.

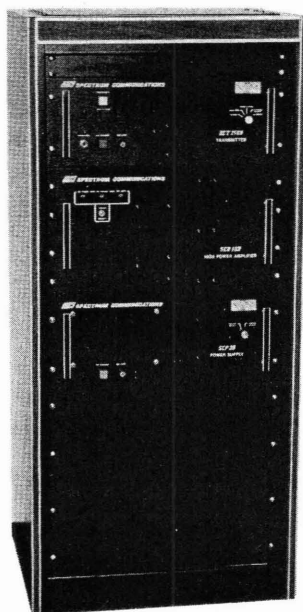
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land mobile radio systems, the vocoder's data rate must be decreased. This rate reduction is accomplished within the vocoder, which removes redundancy in the speech signal waveform.

Therefore, the *improved multiband excitation* (IMBE) vocoder, based on its superior test performance and availability, is selected for use in F-TDMA equipment. The IMBE vocoder's data rate allows two users to communicate simultaneously within the channel capacity of 16kb/s in a 12.5kHz channel.

Linear amplifiers and TDMA

Continuous phase modulation (CPM) is used in the uplink. This method offers a particular benefit in digitally modulating mobiles and portables because non-linear amplifiers still cost less than linear amplifiers. The cost advantage, along with the higher efficiencies of non-linear amplifiers, makes CPM a logical choice for portable F-TDMA radios.

Using DQPSK on the downlink re-

quires a linear or linearized amplifier; still, modulation is used only for base stations where this technology is widely available. DQPSK is a good modulation choice for the downlink because any receiver capable of modulating and demodulating DQPSK is

By narrowing the bandwidth and applying F-TDMA technology, multiple spectrum savings are achieved.

capable of decoding other modulation schemes.

Talk around or simplex

Because F-TDMA is compatible with 12.5kHz operation, talk around is

also possible.

Any F-TDMA radio can operate in the talk-around mode by operating in an FDMA mode, using QPSK modulation and 12.5kHz digital talk around. QPSK is a 4-phase modulation technique in the same family as CPM. Channel efficiency is not compromised when the radio operates in the talk-around mode.

F-TDMA on existing channels

By narrowing the bandwidth and applying F-TDMA technology, multiple spectrum savings are achieved.

The F-TDMA technology provides the equivalent of 12.5kHz channelization on existing 25kHz allocations (e.g., 806MHz-821MHz or 450MHz-470MHz), a twofold increase in current National Public Safety Planning Advisory Committee (NPSPAC) channel allocations (821MHz-824MHz), and a migration path to the proposed 12.5kHz channels below 512MHz.

(continued on page 80)

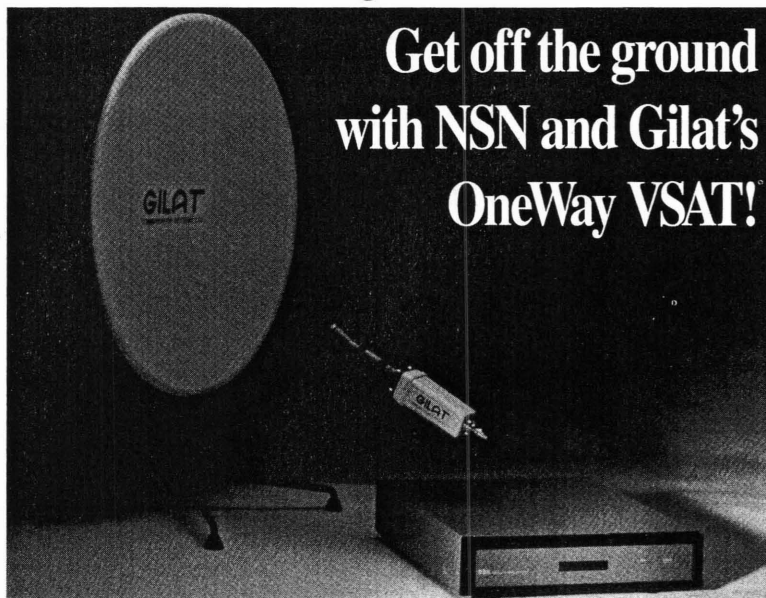
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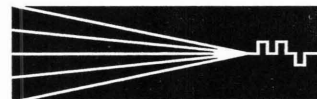
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(continued from page 54)

On-center narrowband channels will be used. This method allows existing channels to be upgraded without having to create new narrowband channels.

F-TDMA on refarmed spectrum

The refarmed spectrum allows two TDMA carriers 12.5kHz apart on a 25kHz channel. (See Figure 3 on page 52.) This method allows four users on the equivalent of one current channel.

F-TDMA technology provides the best solution to satisfy the need for increased spectrum efficiency in land mobile radio while offering the flexibility of supporting high data rates. Moving to narrowband channels alone, as proposed by Project 25, is inadequate to meet the demand for increased spectrum. (Project 25, a joint effort of the Associated Public-Safety Communications Officers, the National Association of State Telecommunications Directors and federal

agencies, seeks to set digital radio standards for public safety radio communications.)

Although the use of analog technology may continue for some time, in due course digital will supplant analog. It is imperative that sound migration plans be available for the transition from analog to digital systems and for maintaining backward compatibility between existing and upgraded systems.

F-TDMA offers backward and forward compatibility. F-TDMA technology has a migration path to achieve a capacity of four users in 25kHz channels under uniform emissions standards. Furthermore, the F-TDMA architecture is fashioned to accept increases in spectrum efficiency.

F-TDMA will benefit significantly from technological advances in other fields. Microprocessors, software and batteries will be leveraged for private land mobile radio systems. Using such components provides economies of scale that will help to lower the

price of TDMA.

F-TDMA data performance

F-TDMA inherently provides 67% higher throughput (16kb/s vs. 9.6kb/s in 12.5kHz channels) than FDMA.

Data applications that require transmissions of fingerprints and retinal scans require wider bandwidths. F-TDMA provides the flexibility to combine bandwidths of multiple time slots and pass data, which requires a wider bandwidth to transmit.

F-TDMA offers a path that allows refarming of existing spectrum below 470MHz to 12.5kHz spacing by on-center channel migration while delivering a 100% increase in efficiency.

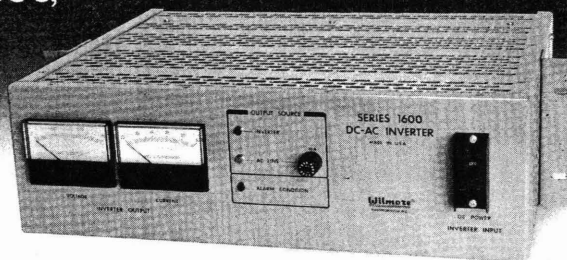
Current 800MHz radio users can maintain their existing 25kHz channels and gain a progressive efficiency increase to four users per channel using F-TDMA.

F-TDMA will be developed and implemented with product availability within the next two years.



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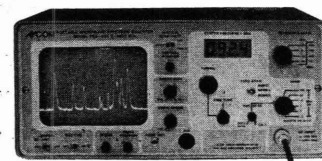
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Solving intermodulation interference

Common problems, real-world examples and solutions

Part 2 of 2

Last month we laid the groundwork for understanding intermodulation interference at co-location sites. This month we'll illustrate some common problems with real-world examples and suggest some practical solutions.

Example 1: Interference from AM station. It may be hard to believe, but an AM station operating at 1040 kHz actually caused interference to a U.S. Postal Service repeater operating in the VHF band. The USPS repeater transmitted on 164.175 MHz and received on 166.250 MHz, and it employed a private line (PL) tone. The symptom resembled a stuck microphone. After the first user keyed his or her portable radio, the repeater remained keyed indefinitely. Users could hear white noise in the background and assumed the repeater was busy and unavailable. In reality, a modest uplink signal would recapture the repeater, but users were not aware of this fact and assumed the repeater was unusable.

An AM broadcast tower, operating on 1040 kHz, was located a few blocks from the post office. A quick look at potential IM products revealed the following: $164.175 + 2(1,040) = 166.255$ MHz, only 5 kHz from the repeater receive frequency. Because the repeater employed good filters and a ferrite isolator, the consulting engineer concluded that the likely source of the IM product was passive IM in the repeater's antenna. The IM product did not create enough distortion to significantly alter the PL tone, so the site receiver always detected a valid tone and kept the repeater keyed. The antenna had recently been replaced to no avail and there was no enthusiasm for a second replacement, so the radio shop programmed different PL tones on the transmit and receive fre-

quencies and the problem disappeared.

Example 2: Channel 23 (TV) interference to GPS L1 channel. The Global Positioning System (GPS) employs two main radio channels, L1 and L2. Civilian GPS receivers rely primarily on the L1 channel, which is a 1 megachip per second spread spectrum signal centered on 1575.47 MHz. A mountaintop site in Colorado is home to both Channel 23 and a nationwide paging company that uses GPS to synchronize its simulcast paging network. The visual carrier for Channel 23 is 525.25 MHz, and the third harmonic of this carrier is 1575.75 MHz, well within the 1 MHz bandwidth of the L1 channel. Immediately after Channel 23 went on the air, the paging station lost synchronization and the company's technician called the consulting engineer for help. A bet-

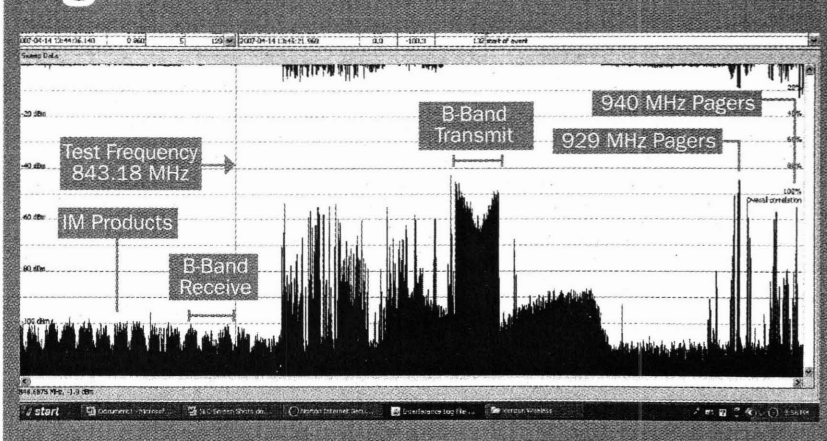
isted. There is not always a single source of IM interference.

The second harmonic of the aural carrier of Channel 66 (TV) falls on 1575.5 MHz, and it also can interfere with GPS L1 receivers.

Example 3: 900 MHz pager interference to the B-band cellular operator. This problem occurred at two different cell sites, and although 900 MHz pagers were involved in both cases, the physical interference mechanism was quite different at each site.

It is fairly well known in the cellular industry that paging transmitters operating in the 929–932 MHz bands can cause interference to the B-band cellular operator. The B-band operator receives in the bands 835–845 MHz, 846.5–849 MHz and transmits in the bands 880–890 MHz and

Figure 1



ter lowpass filter on the Channel 23 transmitter helped somewhat, but the problem persisted because the GPS receiver did not adequately filter the television signal and the third harmonic of Channel 23 also was created in the receiver front end. The receiver was replaced with one that included a better front end filter. Note that in this case, both transmitter-generated harmonics and receiver-generated harmonics ex-

isted. Many B-band operators employ the TIA-95 CDMA air-link standard with 1.25 MHz-wide channels. If one of these CDMA carriers is transmitting on 886.44 MHz and a nearby paging transmitter is operating on 929.5625 MHz, a potential IM product results: $2(886.44) = 1772.88$ MHz, which falls in the reverse link band of the cellular operator. Because the transmit signal

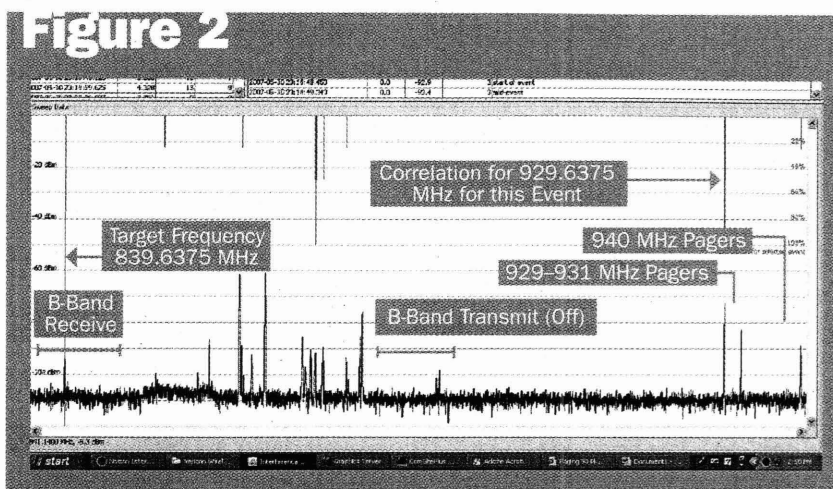
is 1.25 MHz wide, the IM product will be at least 1.25 MHz wide (closer to 2.5 MHz in this case) and at least two CDMA receive channels are affected by this interference.

At the first site, the interference was intermittent and the consulting engineer set up an automated test with a spectrum analyzer connected to the diversity antenna on the affected sector. The control software was set to store each trace whenever energy appeared above a pre-determined threshold on a frequency known to be out-of-service. Figure 1 shows a spectrum analyzer trace for an interference event at this site. Note that the interfering signals are similar to individual CDMA carriers, but roughly twice as wide and with a more rounded shape.

This wider bandwidth and rounded shape is consistent with the second harmonic of a CDMA signal. Note that interferers appear at regular intervals and the interference extends well below the cellular operator's receive band. The reader will also notice a notch exists in the B-band transmit spectrum. This notch is from a notch filter inserted in front of the spectrum analyzer to suppress the transmit signals and thereby prevent the analyzer front end from being driven into saturation.

Further tests showed that the IM products at this site only occurred when the cellular operator was transmitting. After eliminating both transmitter intermodulation and receiver intermodulation as sources, the operator concluded that the cellular antennas were generating passive IM and should be replaced.

The second site is congested with FM broadcast, TV broadcast, VHF repeaters, UHF repeaters, 800 MHz public safety radio, 800 MHz ESMR, 800 MHz cellular, 900 MHz paging and PCS. There are 14 paging transmitters at this site, and measured levels at the affected sector antenna (before filtering) are as high as 0 dBm. The paging antennas are less than 50 feet from the B-Band sector antennas. Unlike the first site, interfer-



ence at this site was still present when the cellular operator was not transmitting. Also, the interference was narrow-band. (See Figure 2.)

An examination of the data showed that the interference always fell exactly 90 MHz below one of three 929 MHz or 931 MHz paging frequencies in use at the site. Further investigation revealed that the paging transmitters employed a 90 MHz intermediate frequency and upconverted the signal from there. Thus, either the local oscillator for the upconversion or an IM product was the source of the interference. The paging transmitters lacked adequate bandpass filters which would have prevented this problem. It is interesting to note that the cellular antennas at the second site were exposed to much higher levels from paging transmitters than at the first site, but they did not create passive IM interference. Thus, the level of PIM rejection differs widely between manufacturers.

Example 4: Image frequency interference to public-safety receiver. Image frequency interference is a form of receiver IM that often is overlooked. Consider a superheterodyne receiver with a first IF of f_{IF} tuned to frequency f_{RF} . The local oscillator (LO) frequency is either $LO = f_{RF} + f_{IF}$ or $LO = f_{RF} - f_{IF}$, depending on the receiver design. An image frequency, f_{IMAGE} , is an undesired frequency that is exactly $2 f_{IF}$ above or below the tuned frequency, f_{RF} , for LO frequencies below or above the tuned

frequency, respectively. For the most part, the mixer does not distinguish between a tuned frequency and an image frequency and it will generate an IF signal from either. The key to eliminating image frequency interference is to employ a good filter at the receiver front end to attenuate the image frequency.

Consider this example: If the tuned frequency is 807.25 and the IF is 45 MHz, the image frequency is $807.25 + 2(45) = 717.25$, which is the visual carrier for Channel 55 (TV). Thus, if the receiver front-end filter does not adequately attenuate 717.25 MHz and the television visual carrier is strong, harmful image frequency interference will occur. Of course, image frequency interference from analog television transmitters operating on Channel 55 will not exist after Feb. 17, 2009, when the digital television transition is complete and all television stations will be digital and will be confined to Channels 2-51 (with a few exceptions). ■

Jay Jacobsmeyer is president of Pericle Communications Co., a consulting engineering firm located in Colorado Springs, Colo. He holds bachelor's and master's degrees in electrical engineering from Virginia Tech and Cornell University, respectively, and has more than 25 years experience as a radio frequency engineer.

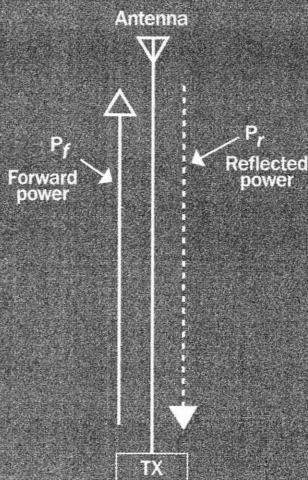
Understanding standing wave ratios

Be careful, as transmission line loss can profoundly impact your calculations

The term standing wave ratio, or SWR—sometimes called voltage standing wave ratio, or VSWR—frequently is a subject of discussions/arguments among communication technicians, hams and others involved in the radio communications field. The easiest way to explain SWR is through an example.

Figure 1

An in-line directional wattmeter is used to measure the forward and reflected power at the transmitter. The results may be quite misleading—especially if the normal matched-line loss of the transmission line is significant. See text for details.



In Figure 1, a transmitter is connected to an antenna through a 100-foot length of transmission line. Let's suppose in this instance that the transmission line is loss-less (though no such circumstance exists in the real world.) Further suppose that the antenna is not properly matched to the 50 ohm system and that 25% of the RF power is reflected back down the line from the antenna toward the transmitter. With the in-line directional

wattmeter connected between the transmitter and the transmission line, the wattmeter indicates 100 W forward power and 25 W reflected power. In Figure 1, P_f represents the forward traveling wave (forward power) and P_r represents the reflected traveling wave (reflected power).

The voltages of the forward and reflected waves will combine at different phase angles along the length of the transmission line. A phasor analysis will reveal that voltage maxima (anti-

nodes) and voltage minima (nodes) exist along the length of the transmission line. Figure 2 shows the representation of the standing wave pattern on the transmission line. The VSWR figure is derived from the ratio of the voltage at the maxima to the voltage at the minima. So all we have to do is calculate the voltage of the forward wave and the voltage of the reflected wave and express the sum and difference of these two voltages as a ratio. This is the standing wave ratio. The calculations are as follows:

First, from the power and impedance we can calculate the voltage using Equation 1. Since the impedance is 50 Ω and the power in the forward wave is 100 W, the voltage of the for-

Equation 1

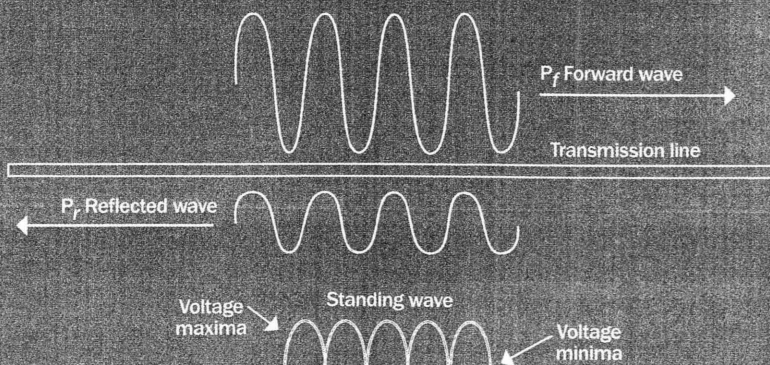
$$V = \sqrt{PZ}$$

Equation 2

$$V = \sqrt{PZ} = \sqrt{100 \times 50} = \sqrt{5000} = 70.71$$

Figure 2

Forward and reflected traveling waves on a transmission line produce a resultant standing wave with voltage maxima and minima. The standing wave ratio is the ratio of the voltage maxima to the voltage minima.



Equation 3

$$VSWR = \left[\frac{70.71 + 35.35}{70.71 - 35.35} \right] = \left[\frac{106.06}{35.35} \right] = 3.00:1$$

Equation 4

$$SWR = \frac{1 + \sqrt{r}}{1 - \sqrt{r}}$$

Equation 5

$$SWR = \frac{1 + \sqrt{.25}}{1 - \sqrt{.25}} = \frac{1 + .5}{1 - .5} = \frac{1.5}{.5} = 3:1$$

ward wave is 70.71 (see Equation 2). The voltage of the reflected wave is calculated in the same manner, and the result is 35.35 V. The VSWR is computed as the ratio of the sum (maxima) to the difference (minima) of the two voltages, or 3:1, as shown in Equation 3.

Most technicians use a directional in-line wattmeter to make forward and reflected power measurements in order to determine transmitter output power and reflected power. With some expe-

rience, the technician can get a quick mental evaluation of the degree of match or mismatch by comparing the forward and reflected power levels. Basically, the technician is comparing the percentage of reflected power to forward power.

In the example illustrated in Figure 1, the forward power is 100 W and the reflected power is 25 W. Thus, the percentage of reflected-to-forward power is 25%. Experienced technicians know that a reflected power of 25% translates to an SWR of 3:1. If we let r represent the ratio of reflected power to the forward power, we can use Equation 4 to calculate the SWR. Using the example in Figure 1, we find that SWR also is 3:1 (see Equation 5).

The term return loss often is used to indicate the degree of match or mismatch instead of the term SWR. To understand return loss, refer back to Figure 1. The forward power is 100 W, and the reflected power is 25 W.

The return loss is simply the expression in decibels of the ratio of the forward power to the reflected power and is calculated by the formula in Equation 6.

Thus, a return loss of 6 dB is equivalent to an SWR of 3:1. It also is possible to convert SWR directly to return loss using Equation 7 on page 60. Table 1 provides conversions between SWR and return loss for several SWR ratios.

In the example of Figure 1, the line loss was theoretically zero—but this never happens in the real world. Let's look back at Figure 1 and see what happens if the transmission line loss is 3 dB at the given length and frequency of operation. This changes things quite a bit. The power output from the transmitter (P_f) is still 100 W. However, since the line loss is 3 dB, the forward power arriving at the antenna is only 50 W (down 3 dB). Now the mismatch at the antenna causes 25% of the forward power to be reflected back toward the transmitter. Thus, 12.5 W of power

Continued on page 60

Table 1

Return Loss (dB)	VSWR
10	1.92
12	1.67
14	1.5
16	1.37
18	1.28
20	1.2

Equation 6

$$RL = 10 \log \frac{P_f}{P_r} = 10 \log \frac{100}{25} = 10 \log(4) = 10(0.602) = 6.02 \text{ dB}$$

Equation 7

$$RL = 10 \log \left[\frac{1}{\left(\frac{S-1}{S+1} \right)^2} \right]$$

Where RL is return loss and S is SWR

Equation 8

$$L_A = 10 \log \left[\frac{1-r}{1-r \left(10^{\frac{L}{5}} \right)} \right]$$

Where L_A is additional line loss, r is reflected/forward power ratio at transmitter and L is normal matched-line loss of transmission line

Continued from page 38

is reflected back down the line. In traveling back down the line, this reflected power is attenuated by 3 dB. This means that the reflected power measured at the transmitter is only 6.25 W.

Using all of this information, we can analyze the results in the following manner. First, the dif-

ference between the forward and reflected power readings at the transmitter is 100 W minus 6.25 W for a net power of 93.75 W. The SWR at the transmitter is 1.67:1, and the SWR at the antenna is still 3:1. Thus, the loss of the transmission line makes the SWR measured at the transmitter look much better than it really is at the antenna.

Remember, too, that the actual

power delivered to the load is the difference between the forward power and reflected power readings on the wattmeter. Now, at the transmitter, the difference between the two readings is 93.75 W, while at the antenna it is 37.5 W.

Because the line loss was 3 dB (representing a 50% loss), the net power delivered to the antenna should be 50% of the net power delivered into the transmission line at the transmitter. Calculations indicate that the net power delivered to the antenna is only 40% of the net power input to the transmission line. Thus, the loss in the transmission line is 60%, or approximately 4 dB. The normal (matched-line) loss of the transmission line is 3 dB. The additional 1 dB of loss is caused by the mismatch at the antenna. Equation 8 can be used to determine the additional line loss caused by a mismatch at the antenna.

When you measure the SWR (reflected and forward power readings) at the transmitter, remember that the line loss between the transmitter and antenna will mask the true SWR that exists at the antenna. The greater the transmission line loss, the greater this masking effect.

In fact, if the normal matched-line loss of the transmission line is 3 dB, the maximum SWR seen at the transmitter would be just slightly more than 3:1—even for a worst-case mismatch at the antenna. SWR increases the line loss above the nominal matched-line conditions. When choosing a transmission line, use the best (read minimum loss) cable that your budget and other practical considerations will allow.

Until next time—stay tuned! ■

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Well-grounded principles

Follow these guidelines to eliminate lightning damage

Few disciplines are dominated by as many myths and wives tales as electrical grounding. Everyone agrees that grounding is important to reduce the occurrence of lightning damage but, paradoxically, few sites have good grounds. This article describes some principles of good grounding.

There are three important reasons for proper grounding: ensuring personnel safety, minimizing radio frequency interference and preventing equipment damage from lightning. This article focuses exclusively on the lightning problem. Lightning is an impulse that can create instantaneous down conductor currents greater than 20,000 amperes. Because it is an impulse, a lightning surge produces alternating current with frequency content as high as 100 MHz.

A good ground is a low-impedance bond to an effective earth electrode system, which is a network of buried conductors around the tower designed to create a low resistance-to-earth. Tower earth electrode systems typically consist of a set of driven ground rods configured in a ring around the tower, with

Equation 1

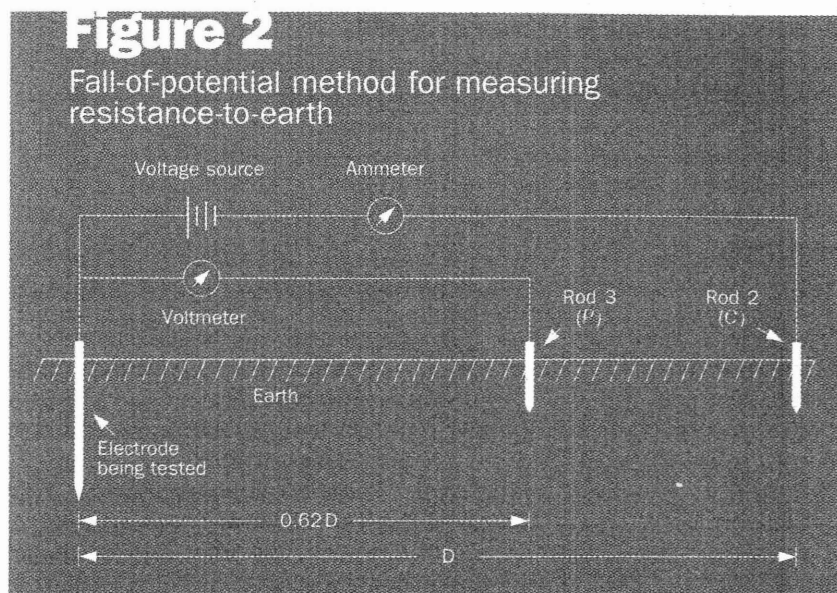
$$R_0 = \frac{\rho}{2\pi l} \ln \left(\frac{4l}{d} \right)$$

is 167 ohms. Typically, one wants a resistance-to-earth less than 10 ohms, so some improvement is needed.

Improvements come in several forms. One can lengthen the rod, increase the diameter of the rod, add more rods, bond to structural steel or other buried metal, or use chemical rods. Chemical rods increase the conductivity of the soil around the rod, increasing its effective diameter.

Real-world grounding systems often are quite complex when dealing with non-homogeneous soil conditions. Measurements are required to verify resistance-to-earth. Unfortunately, one cannot measure resistance-to-earth with a common multimeter. Specialized equipment and techniques are required. The preferred technique is the fall-of-potential method, illustrated in Figure 2.

The resistance-to-earth of the electrode under test is simply V/I , where V is the voltage reading at the voltmeter



and I is the current reading at the ammeter. The voltage source is usually AC, operating at a frequency that is not a harmonic of 60 Hz.

The correct reading is found when distance P is 62% of C [2]. It also is important that C be large relative to the largest dimension—the diagonal—of

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the earth-electrode system. A good rule of thumb is that the ratio of C to the system diameter should be at least 9 to 1.

Equally important as the earth electrode system is the down conductor bonded to it. Lightning creates alternating current in down conductors, so we are interested in the properties of down conductors at high frequencies. We recall that impedance, Z , consists of two parts, as shown in Equation 2.

We know that coils of wire have high inductance, but even a straight wire has some inductance. For example, a straight, 30-foot-long, No. 2 copper wire has an impedance of 9050 ohms—at $f=100$ MHz—almost entirely due to inductance. To reduce inductance, we should avoid sharp bends, but we also can reduce inductance by creating more surface area. For example, a 3-inch-wide copper strap of the same length has an impedance of only 6900 ohms. Strap generally is a better down con-

ductor than wire, but mechanical considerations often dictate that round conductors be used.

Let's wrap up with some common grounding mistakes:

Installing down conductors from high on the tower: Installers are sometimes told they must install a copper down conductor from the antenna to the earth electrode system. The author has seen more than 200 feet of AWG No. 2 wire installed on a tower to meet this misguided requirement. True, copper has better conductivity than galvanized steel, but the impedance of the tower is much lower than No. 2 copper

wire for two reasons: the larger surface area of tower members creates lower inductance and the multiple current paths to the ground create lower overall impedance.

Figure 3 depicts a similar mistake. Here, the installer mounted a bus bar about 10 feet above grade near the base of the tower. He bonded three down conductors to the bus bar, which is good, but he failed to bond the bus bar to a tower member. The large surface area and multiple paths created by tower members create a lower impedance path to ground than the three small down conductors. The concern here is that a lightning surge might find the coaxial cable to be a lower impedance path to ground (through the radio cabinet) than the three down conductors. A mitigating circumstance is that the bend in the coaxial cable creates higher impedance, making it somewhat less attractive to the lightning surge.

Failure to bond building structural steel to an earth electrode system: Copper and steel conductors embedded in concrete also create an effective ground. Such a ground system is called a Ufer ground (after its inventor, Herbert Ufer, who was a consultant to the U.S. Army during World War II). Too often, the designer fails to bring conductors outside the foundation for bonding to the earth electrode system.

When structural steel is bonded to the earth electrode system, the results can be quite dramatic. The author participated in one mountain-top project where the measured resistance-to-earth of a 20-rod, building earth electrode system—prior to bonding to foundation steel—was 150 ohms. After bond-

Equation 2

$$Z = R + j2\pi fL$$

Where R is the resistance of the conductor (which is a function of frequency due to the skin effect), f is the frequency, and L is the inductance of the down conductor.

Figure 3

Failure to bond bus bar directly to tower steel



Multiple down conductors are good, but why is the bus bar not bonded directly to the tower?

ing foundation steel to the earth electrode system (all below grade), the resistance-to-earth was 1.0 ohm.

A word of caution: A Ufer ground consisting solely of the tower foundation is a bad idea. Lightning surges passing through the foundation can vaporize water in the concrete and damage the foundation through rapid expansion of steam. An example of this unfortunate event is shown in Figure 4. This particular tower had no earth electrode system, other than the leg foundations.

Failure to prevent conduit from acting as an RF choke: It is common for building codes to require that in-building, insulated (green wire) ground conductors be routed inside to the building ground through steel conduit. Unless the wire is electrically bonded to the conduit when it enters and leaves it, the steel conduit acts as an RF choke, presenting high impedance to alternating

current, including lightning surges.

Space does not permit us to cover all the best practices for grounding, so the author recommends the industry and government guides listed below. ■

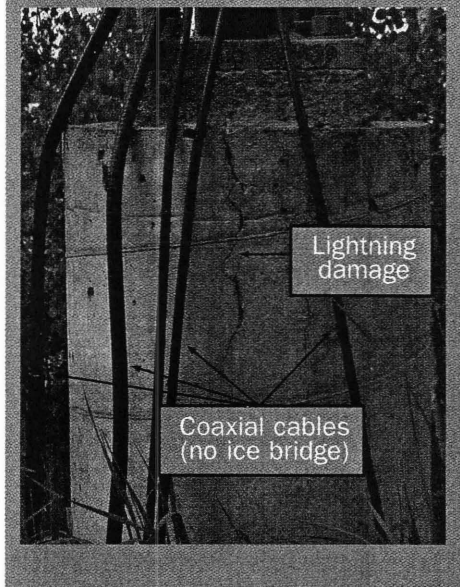
Jay Jacobsmeyer is president of Pericle Communications Co., a consulting engineering firm located in Colorado Springs, Colo. He holds bachelor's and master's degrees in Electrical Engineering from Virginia Tech and Cornell University, respectively, and has more than 25 years experience as a radio frequency engineer.

References:

- [1] Motorola R-56, "Standards and Guidelines for Communications Sites," 2000.
- [2] MIL-HDBK 419, "Grounding, Bonding and Shielding for Electronic Equipment and Facilities," 1987.

Figure 4

Damage caused by lightning surges through concrete foundation

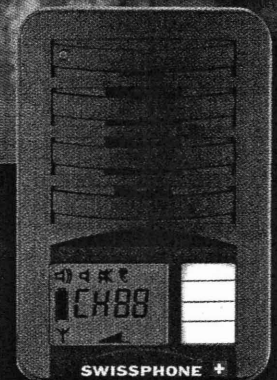


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RF transmission line tips

By Harold Kinley, C.E.T.

There are a few techniques that all technicians should know in order to work and deal with RF transmission lines more effectively. Some myths have been perpetuated over the years. Some are taken as fact without question. The following tips should help to clear up some of the murky water surrounding RF transmission lines.

Trim the coax!

I remember, back during the heyday of CB radios, hearing some of the "experts" on CB telling their friends how to reduce the VSWR on their rig. Almost invariably the advice was "Trim the coax!" Seldom did I hear "Tune the antenna!" If the antenna is properly tuned it will present a 50Ω impedance to the transmission line. If the transmission has the proper 50Ω characteristic impedance, the length of the line will not matter as long as the antenna is properly tuned.

Non-resonant lines

A transmission line is *non-resonant* when it is terminated in an impedance that is equal to the characteristic impedance of the line. Changing the length of non-resonant transmission lines does not change the impedance at the input. It can be 50 feet or 100 feet, but the input impedance will remain unchanged. The only degrading factor will be the loss of the line as the length increases.

Resonant lines

When a transmission line is terminated by an impedance that is *not* equal to the characteristic impedance of the line, it is said to be *resonant*. The length of a resonant transmission line is critical. Changing the length of the line will change the impedance of the line, looking from the source. When an antenna impedance is other than the characteristic impedance of the transmission line, the line length can be adjusted to present a better matching impedance at the output of the transmitter.

Line length vs. impedance

This discussion assumes a 50Ω system impedance and thus the use of 50Ω transmission line—that is, transmission line with a characteristic impedance of 50Ω.

In Figure 1 above, a transmission line is terminated into a 50Ω load. The impedance $1/8\lambda$ from the load is 50Ω. The impedance $1/4\lambda$ from the load is 50Ω. The impedance $3/8\lambda$ from the load is 50Ω. The

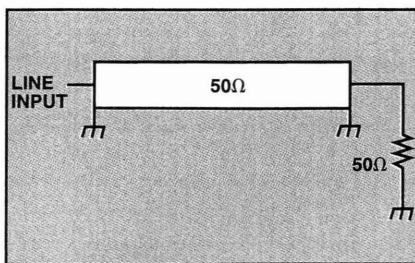


Figure 1. A transmission line terminated into a load equal to the characteristic impedance of the line—50Ω. The impedance will be the same at all points along the line.

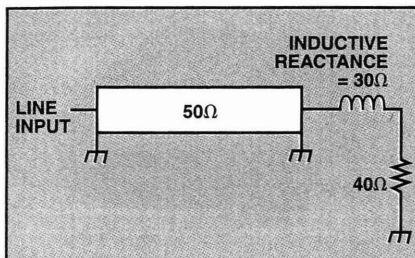


Figure 2. A 50Ω transmission line terminated by a load with the complex impedance of $40 + j30$. The impedance will vary along the length of the line and will repeat at every $1/2\lambda$.

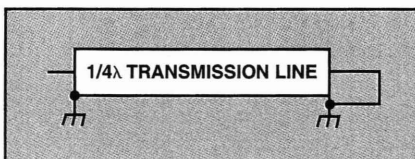


Figure 3. A $1/4\lambda$ transmission line shorted on one end produces an open circuit on the other end.

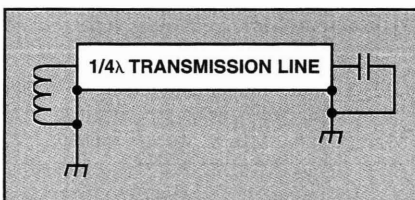


Figure 4. A $1/4\lambda$ transmission line can be used to transform an inductive reactance at one end of the line into a capacitive reactance at the other end of the line.

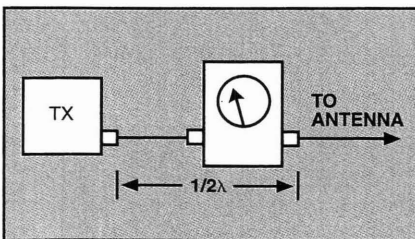


Figure 5. If a transmitter must be tuned into a resonant line, be sure to use a $1/2\lambda$ cable (including wattmeter section). Include velocity factor in cable length calculations.

impedance $1/2\lambda$ from the load is 50Ω.

Thus, the impedance anywhere along the length of a transmission line will remain the same as long as the line is terminated into a load equal to its characteristic impedance. To further qualify this statement, the load must be a *purely resistive load* equal to the characteristic impedance of the line. This means that the load must not contain a reactive component. Written in complex form, the load must be equal to $50 \pm j0$.

To illustrate the preceding point, the formula for impedance is:

$$Z = \sqrt{R^2 + X^2}$$

Now, if the load impedance in the complex form is $40 + j30$, then the resultant impedance is 50Ω. (See Figure 2 at the left.) However, because a reactive component is in the load, a mismatch will result. The actual VSWR at the load will be 2:1. The same VSWR will be seen at all points along the transmission line, assuming a lossless line. However, the actual impedance along the line will vary, and will repeat every $1/2\lambda$.

At a distance of $1/8\lambda$ from the load, the impedance is $100 \pm j0$. At $1/4\lambda$, the impedance is $40 - j30$. At $3/8\lambda$, the impedance is $25 \pm j0$. And, again, at $1/2\lambda$, the impedance repeats the load impedance— $40 + j30$. The cycle repeats every $1/2\lambda$ on the line moving away from the load and toward the generator.

Impedance transformation

From the preceding information, it follows that a transmission line can be used to transform impedance. In fact, a $1/4\lambda$ line can transform an open circuit at one end into a short circuit at the other end. Or, it can transform a pure inductive reactance into a pure capacitive reactance. (See Figures 3 and 4 at the left.) Open or shorted stubs are used to match impedances using the Smith chart. (Smith is a registered trademark of Analog Instruments, P.O. Box 808, New Providence, NJ 07974.)

Normally, any mismatch at the antenna

(continued on page 33)

Kinley, a certified electronics technician, is regional communications manager, South Carolina Forestry Commission, Spartanburg, SC. He is a member of the Radio Club of America. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, which is available for direct purchase. Write to 204 Tanglewyld Drive, Spartanburg, SC 29301. Kinley's email address is hkinley@aol.com.

(continued from page 8)

should be corrected so that the antenna presents a proper 50Ω impedance to the transmission line, which, in turn, will present the 50Ω impedance to the transmitter. Because the transmitter is designed to work into a 50Ω load impedance for optimum performance, every effort should be made to make the impedance at the transmitter equal to 50Ω .

If this is not possible, and the transmitter must be operated into a line with moderate VSWR, the following technique should be used. An in-line wattmeter is connected between the transmitter and the transmission line using a length of coax, such that the length of coax plus the in-line section of the wattmeter are exactly $1/2\lambda$ at the operating frequency. Be sure to take into account the velocity factor of the connecting line. (See Figure 5 on page 8.) With the wattmeter connected through a $1/2\lambda$ section of line (plus wattmeter section), the transmitter is loaded and tuned for proper power output. Then, the wattmeter and connecting line are removed, and the transmission line is reconnected. The transmitter output should see the same impedance it saw during the tuning procedure.

Line loss masks true VSWR

When you check the forward and reflected power at the transmitter output, you are not getting the actual reflected and forward readings that exist at the antenna, unless the line loss between the transmitter and antenna is quite small. For example, suppose that a catastrophic failure occurs at the antenna transmission line connection—either open or short circuit. Theoretically, this represents a VSWR of infinity. However, if the line has a loss of 3dB, the actual VSWR seen at the transmitter will be about 3:1. Obviously, high VSWR levels can go virtually undetected by measurements made at the transmitter when line loss is high. The line loss acts like a pad to smooth impedance bumps. Such padding is desirable on low-level signal generators but not on high-power transmitters.

Lightning protection?

An old practice among many technicians is to use a shorted transmission line stub to provide a degree of lightning protection. A shorted $1/4\lambda$ stub is placed on the transmission line using a "tee" connector. The length of the stub is $1/4\lambda$ at the operating frequency with the velocity factor of the line included. The stub has negligible effect on the system at the operating frequency but puts a dc short

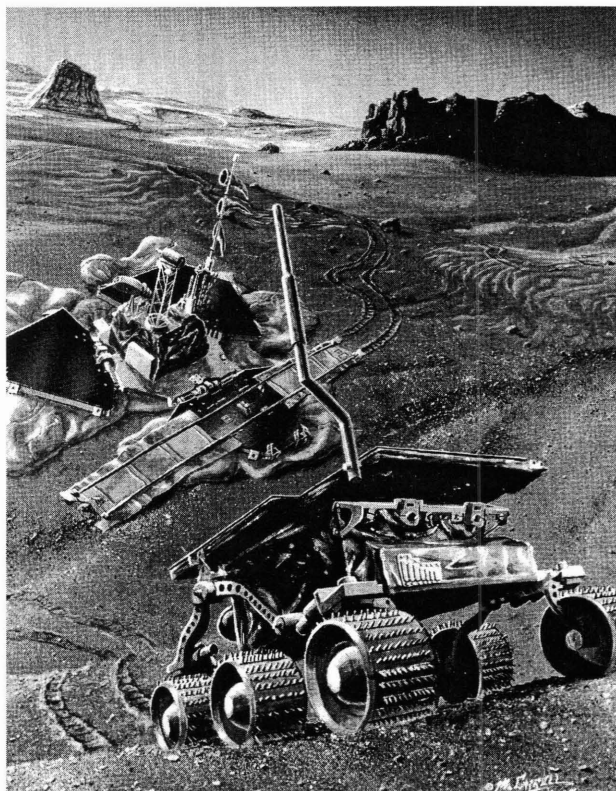
across the line to offer some measure of protection against lightning. I won't argue for or against the merits of such a lightning protection scheme.

Similarly, $1/4\lambda$ stubs can be used as notch filters to notch out interfering signals. Good-quality cables with high velocity factors are generally more effective for such purposes.

I hope this has provided some insight

into working with transmission lines and avoiding some pitfalls along the way. The RF transmission line is more than just a conduit through which RF energy is passed. It is a complex electronic circuit with many features that can be used to great advantage.

Until next time, *stay tuned!*



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How to link mobiles to form a repeater

Simple repeater configurations in Alaska support radio communications links as long as 60 miles. Here's how to build a link for use with two Motorola Maxtrac radios. Similar links can be designed for other units.

By Carlton L. Tedrick, P.E.

During a consulting assignment in Alaska, a simple, reliable and effective method of linking two mobile, underdash transceivers in a back-to-back repeater configuration was needed. Commercially made devices are available to accomplish the task. But because they are designed to be universal, they tend to be complex, and many are difficult to adapt to specific needs. The immediate requirement was for a method

Tedrick, a consulting engineer, owns Communications Service and Engineering, Nacogdoches, TX. He also is an independent contractor with Lockard & White consulting telecommunications engineers, Houston. This information first appeared in Lockard & White's *Network*.

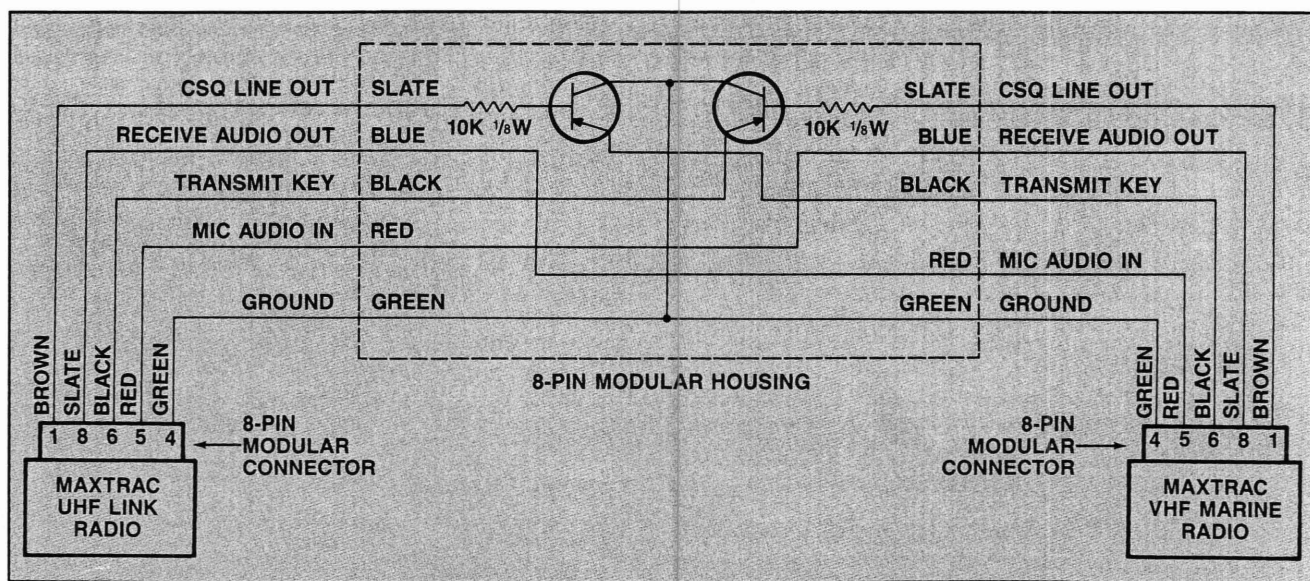
of reusing a number of Motorola Maxtrac mobile units.

The first task was to identify a control point in each receiver that would provide a change of logic state when a signal opened the squelch circuit. Such a point was found at pin 2 of J6 in the Maxtrac. A small jumper was run from that point to the unused pin 1 of J11 on the front panel. This jumper provides a change of state from 4Vdc to 0Vdc when the squelch opens. A small, general purpose transistor completes the transmitter keying circuit.

Locating an adequate point from which to derive received audio to drive the microphone input of the adjacent radio was next. The transmitter's input sensitivity is approximately 80mV rms for 60% carrier deviation. An oscillo-

scope was used to look for suitable fixed audio output points in the receiver. *Fixed* is the key word because any *variable* point, such as the volume control center tap, invites trouble. The signal level at the high side of the volume control was too high, and the handset output produced a level that was much too low. A 1,800Ω, 1/8W resistor applied between pins 1 and 4 of J8 provided just the right level of audio to feed the microphone input from the handset audio connection of J11.

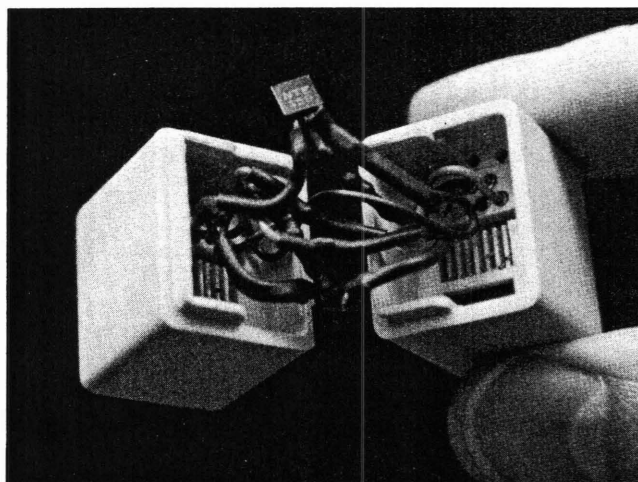
This audio point is *after* the demphasis circuitry, as it should be to drive the microphone input circuitry of the adjacent transmitter. Once the linkup was complete and set up for bench testing, the transmitter deviation was set to produce a "one-to-one" ratio—that is,



This diagram shows how to wire the double-ended female 8-pin RJ modular connector and the RJ connectors that plug into the Maxtrac radios.



RJ-type modular connectors are used in the Maxtrac radios, so a double-ended female 8-pin RJ modular connector device was chosen to house the link. This forces compact wiring, and short leads are required in such configurations to reduce the possibility of stray RF pickup.



The double-ended female RJ connector is snapped apart carefully and the two transistors and cross-connections are soldered in place. Heat-shrink tubing was used on each connection because of small working quarters.

a received deviation of 3kHz produced a deviation of 3kHz on the opposite transmitter. Take into consideration continuous-tone controlled squelch system (CTCSS) signaling and its effect on overall deviation settings when making these adjustments.

These modifications were performed on both radios to be used in the linked configuration. They are minimal modifications. Nevertheless, they allow the Maxtrac radios to continue to be used as conventional mobile units, in most instances. These modifications, though, will cause increased handset audio when a handset is used.

Development of a *simple and reliable* "link" device to connect the two radios was necessary. RJ-type modular connectors are used in the Maxtrac radios. A double-ended female 8-pin RJ modular connector device was chosen to house the link. This forces compact wiring, and short leads are required in such configurations to reduce the possibility of stray RF pickup. This type of wiring permits ordinary flat telephone cable and crimp-on RJ connectors to be used, creating a virtually foolproof interconnection arrangement. Remember: *Keep all leads short.*

The double-ended female RJ connector is snapped apart carefully and the two transistors and cross-connections are soldered in place. Heat-shrink tubing was used on each connection because of small working quarters.

All that remains is to build two *extremely short* 8-conductor cables with 8-pin male RJ connectors crimped onto the ends. These cables should be assembled with the connectors in the standard configuration, with locking tabs on all connectors facing in the same direction. This assures proper interconnection and polarity.

With construction complete, bench-testing is in order. Here is where the foolproof nature of the design becomes apparent. You can connect the link device to the radios without regard to orientation or polarity—it should work fine. Remember, though, this particu-

With construction complete, benchwork testing is in order. Here is where the foolproof nature of the design becomes apparent.

lar device is designed to work with *two identical radios* (Maxtracs in this case), and probably will not work with different radios. A similar arrangement probably can be developed for similar radios from other manufacturers.

One minor drawback should be not-

ed. The CSQ line used to control the adjacent transmitter push-to-talk (PTT) is at a logic high when the squelch is closed (no signal received). The CSQ line goes to logic low when the squelch opens (signal received). If one of the radios fails or loses power, the same condition occurs—logic high goes to logic low. This will key the PTT lead of the adjacent transmitter, resulting in an unmodulated carrier on the air.

To prevent this from becoming a problem, the Maxtrac's time-out interval should be programmed for less than three minutes. Then, if one radio fails or loses power, the "dead" carrier will be limited to the timer interval. The transmitter then will shut down until the problem is corrected.

Circuitry to prevent the PTT line from being keyed when one radio fails or loses power can be designed, but it would complicate the link arrangement and would require more extensive modification of the radios. The simplicity of the system as described outweighs the minor drawback.

Several of these linked radio pairs are deployed in Prince William Sound and the Gulf of Alaska at unattended, solar-powered mountaintop sites. They appear to perform as intended. They do not exhibit any sensitivity to nearby RF fields from other radios. This is an inexpensive solution to a unique problem, one that may be useful to others.



Verifying talk-back coverage

Drive tests can be used for more than just talk-out performance

Last month we talked about methods for conducting drive-test surveys with emphasis on talk-out, or downlink measurements. We neglected to mention talk-back, or uplink performance. The purpose of this column is to examine two methods for measuring uplink performance: (1) direct measurement of the uplink and (2) estimating uplink performance from downlink measurements.

Let's first consider direct measurement. Although not common, direct measurement of the uplink is feasible today using modern computer-controlled test receivers and GPS. A direct uplink measurement requires a test transmitter in the drive-test vehicle. For convenience, the same drive-test vehicle can serve as the uplink transmitter and the downlink receiver, but one must be careful to not de-

dense the downlink receiver with the uplink transmitter. A test receiver must be installed at each repeater site to continuously collect and record measurements. For large networks, the number of required test receivers may make the direct uplink test impractical.

Unlike the downlink drive-test survey, vehicle location is not available at the receiver—it is available at the transmitter. Thus, measurements must be post-processed and time-synchronized using GPS time from each test station, so latitude and longitude can be applied correctly to each receiver measurement.

The observant reader will recall that each measurement is actually a linear

average of at least 50 subsamples over a distance of at least 40 wavelengths. It is unlikely that the receiver will know in real time when the transmitter has traveled 40 wavelengths, so the subsamples must be stored locally and averaging is done as part of post-processing. Thus, the storage requirements for the uplink test are more demanding than they are in the downlink drive-test survey.

Furthermore, we must remember that each receiver tells us the performance at just one repeater site. If the network employs voting receivers, further processing must be done to simulate the voting algorithm and pick the strongest signal. (Voters actually select the best receiver based on some measure of signal-to-noise ratio, not signal strength, so this is an approximation.)

ing are completely uncorrelated because of frequency separation (except for simplex systems), the local mean amplitude of the uplink signal is nearly identical to the downlink signal (for the same effective radiated power, or ERP) for the most widely used bands [1].

So uplink performance follows downlink performance. But we're not done yet. There are still important differences between the uplink and downlink paths. The downlink transmitter typically operates with more power than the uplink transmitter (e.g., 100 W versus 3 W) and the uplink receiver system typically has better sensitivity than the downlink receiver because of the tower-top amplifier (e.g., -126 dBm at the antenna port versus -118 dBm).

Equation 1

$$L_{max} \text{ (dB)} = \text{ERP (dBm)} + 2.2 + G_r \text{ (dBd)} + 2.2 - P_{th} \text{ (dBm)}$$

where **ERP** is the effective radiated power, the factor 2.2 dB is the gain of a half-wavelength dipole, **G_r** is the gain of the receive antenna, and **P_{th}** is the sensitivity of the receiver.

At this point, we might be a bit discouraged. So is there an alternative? Yes, we can estimate uplink performance from downlink measurements if we employ two important principles:

■ **Antenna reciprocity:** The principle of antenna reciprocity guarantees that the antenna pattern is the same whether the antenna is receiving or transmitting.

■ **Local mean** is nearly constant between uplink and downlink. Although instantaneous measurements of the uplink and downlink in Rayleigh fading

are completely uncorrelated because of frequency separation (except for simplex systems), the local mean amplitude of the uplink signal is nearly identical to the downlink signal (for the same effective radiated power, or ERP) for the most widely used bands [1]. So uplink performance follows downlink performance. But we're not done yet. There are still important differences between the uplink and downlink paths. The downlink transmitter typically operates with more power than the uplink transmitter (e.g., 100 W versus 3 W) and the uplink receiver system typically has better sensitivity than the downlink receiver because of the tower-top amplifier (e.g., -126 dBm at the antenna port versus -118 dBm).

What we need at this point is a performance metric common to the uplink and the downlink. We'll use this metric to scale the downlink measurements to quantify uplink performance. The metric of choice is the maximum path loss, or L_{max} , which is the path loss that results in a received signal exactly equal to the receiver's static sensitivity. For example, the 12 dB SINAD sensitivity of a typical 800 MHz portable radio is -118 dBm. If the ERP of the repeater site is 57 dBm (500 W) and the receive antenna gain

is 0 dBd, then the maximum path loss, using Equation 1, is 179.2 dB.

Now, let's compute the maximum path loss on the uplink. We'll assume the receiver system sensitivity referenced to the antenna port is -126 dBm and the portable radio operates with an ERP of 35 dBm (3 W). The receive antenna gain, G_r , is 10 dBd. We'll also assume that the repeater site uses a duplexed antenna or equivalently, that the transmit and receive antennas are identical and at the same tower height. Using Equation 1, we find that the maximum path loss for the uplink is 175.2 dB.

The link imbalance is the difference between the downlink maximum path loss and uplink maximum path loss and can be found by applying Equation 2. A positive value of link imbalance indicates a system that favors the

downlink, while a negative value indicates a system that favors the uplink. A perfectly balanced system has a link imbalance of 0 dB.

For our duplexed antenna case, the link imbalance is $\Delta = 179.2 - 175.2 = 4$ dB. To estimate uplink coverage for this case, we should scale each of our downlink measurements downward by 4 dB.

Typically, downlink performance is measured by computing the service area reliability for a particular service threshold, say -101 dBm. The service threshold is only valid for a certain value of receiver sensitivity, which, as

we have noted, is quite different for the uplink and the downlink. Thus, when we scale the downlink measurement by 4 dB, we are not saying that the signal at the repeater site is 4 dB less than at the portable. Instead, we are saying that the signal is 4 dB closer to the receiver noise floor.

An imbalance favoring the downlink is typical of land mobile radio systems, but system designers are aware of this imbalance, and they usually apply the following techniques to help correct it:

■ higher gain receive antenna at the repeater site,

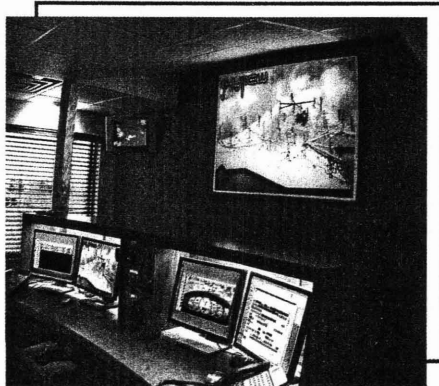
Equation 2

$$\Delta = L_{\max}(\text{downlink}) - L_{\max}(\text{uplink})$$



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We cannot simply scale downlink drive test measurements to estimate uplink performance. Instead, we need drive-test measurements from the omnidirectional antennas.

- antenna diversity,
- voting receivers.

A higher-gain receive antenna is the simplest solution, but it will have a narrower vertical beamwidth and will therefore perform poorly near the repeater site. There also are practical limitations to antenna gain. For an omnidirectional 800 MHz antennas,

for example, most manufacturers top out at 12 dBd.

Antenna diversity almost always is used in cellular networks, but it's rarely used in public-safety radio networks.

Conversely, voting receivers are commonly used in public-safety networks, but rarely used in cellular networks. Note that a voting receiver sys-

tem actually is a form of antenna diversity, with the antennas separated by miles rather than feet.

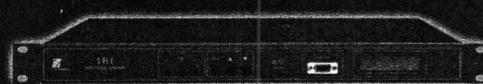
So how do we estimate uplink performance when one or more of these techniques are used? To answer this question, let's consider a second example where voting receivers are used to help correct the link imbalance. In this case, the network consists of three simulcast repeater sites with directional transmit antennas, omnidirectional receive antennas and voting receivers (see Figure 1 on page 70).

We can no longer apply antenna reciprocity directly because reciprocity only applies if the same antenna is used for transmit and receive. Thus, we cannot simply scale downlink drive-test measurements to estimate uplink performance. Instead, we need drive-test

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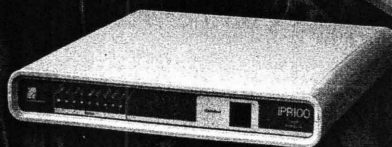


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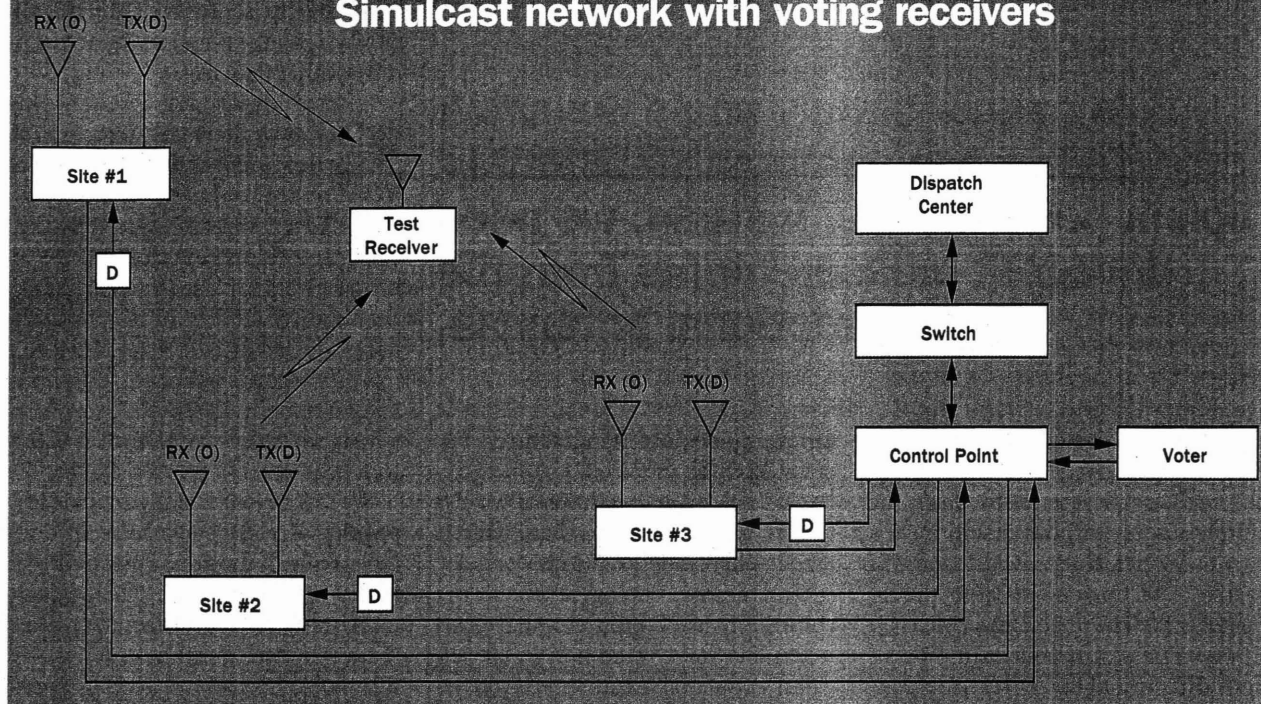
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Figure 1**Simulcast network with voting receivers**

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measurements from the omnidirectional antennas.

But how do we collect measurements from omnidirectional receive antennas? We could revert back to the direct uplink measurement approach with all of its complexity and cost or we could consider an alternative: Transmit a test signal from an omnidirectional antenna.

If the system does not employ tower-top amplifiers, one option is to temporarily install a duplexer on the receive antenna line and connect a test transmitter to the duplexer. The duplexer prevents the test transmitter from interfering with the site receivers while introducing some insertion loss.

If tower-top amplifiers are used, an alternative approach is to temporarily install an omnidirectional test antenna on the tower and connect the test transmitter to it. With test transmitters

connected to each omnidirectional antenna, we collect downlink measurements in the conventional way and scale them for uplink performance as described above. ■

Jay Jacobsmeyer is president of Pericle Communications Co., a consulting engineering firm located in Colorado Springs, Colo. He holds bachelor's and master's degrees in Electrical Engineering from

Virginia Tech and Cornell University, respectively, and has more than 20 years experience as a radio frequency engineer.

References:

- [1] W.C. Jakes, ed., *Microwave Mobile Communications*, IEEE Press Reissue, 1994.
- [2] J.M. Jacobsmeyer and G.W. Weimer, "Guidelines for Conducting Drive Test Surveys for 800 MHz Rebanding," October 1, 2005. Available at www.pericle.com.

Correction:

In the April 2006 issue of *MRT*, Equation 2 in Tech Speak was incorrect. The correct equation appears below. *MRT* regrets the error.

Equation 2

$$n = \frac{Z_{\alpha/2}^2 p(1-p)}{d^2}$$

where $Z_{\alpha/2}$ is the argument of the unit normal distribution for a confidence of $1-\alpha$ and d is one-half of the confidence interval. For example, for 90% confidence, $Z_{\alpha/2} = 1.65$.

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Working with RF transmission lines

By Harold Kinley

As stated in January's "Technically Speaking," when a transmission line is terminated in its characteristic impedance, the line is *non-resonant*. The length of a non-resonant line is not important, except when considering the attenuation of the desired signal between the transmitter and the antenna, or between the antenna and the

receiver. The impedance anywhere along the line will be the same.

Standard system impedance

In the land mobile radio field the standard impedance is 50Ω. Mobile radio transmitters and power amplifiers are designed to work into a 50Ω load impedance. Antennas are designed to present a 50Ω impedance to the transmission line. The coaxial cable that de-

livers power from the transmitter or amplifier to the antenna has a characteristic impedance of 50Ω.

In Figure 1 below, a 50Ω generator is connected to a 50Ω transmission line that is terminated in a load. If the impedance of the load is 50Ω (purely resistive), all of the power delivered to the load will be dissipated in the load. If the load impedance is anything other than 50Ω (purely resistive), power will be reflected back down the transmission line toward the source or generator.

If the impedance mismatch is a worst-case mismatch (open or short circuit), then all of the power delivered to the load will be reflected back down the line toward the source or generator. Thus, two traveling waves will be on the transmission line: a *forward-traveling* wave and a *reflected-traveling* wave. The forward traveling wave is often referred to as the *incident* wave. The phase relationship of the incident and reflected waves will vary along the length of the line.

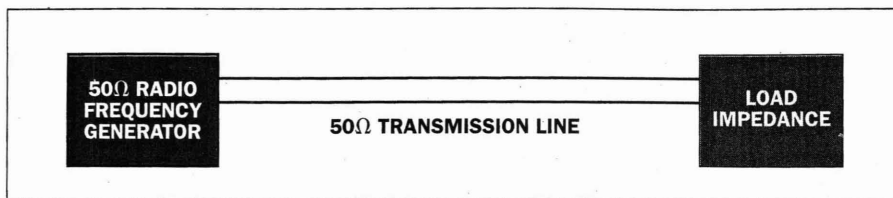


Figure 1. A 50Ω signal generator is connected to a load impedance through a 50Ω transmission line. The load impedance can be anything from a short circuit to an open circuit and everything between those two extremes.

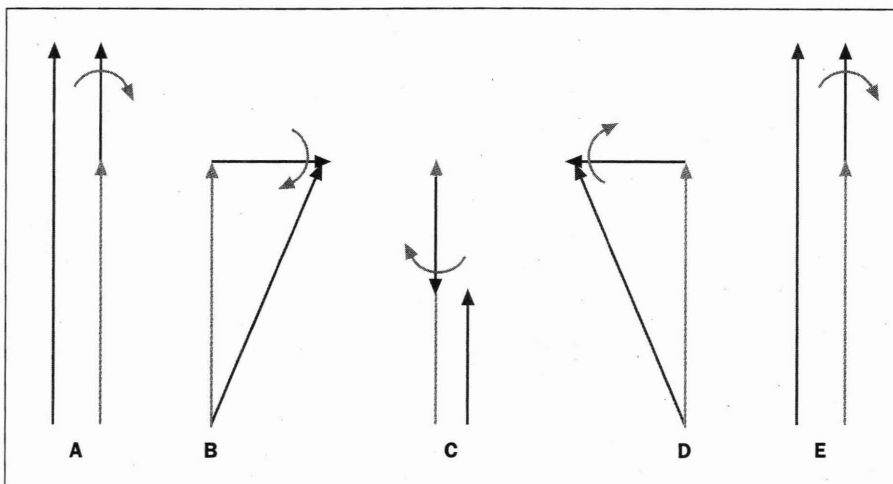


Figure 2. These phasors indicate how the standing wave is constructed. The green phasor represents the voltage of the incident wave, while the red phasor represents the voltage of the reflected wave. The vector *resultant* (blue) represents the voltage of the standing wave. Because the green phasor is the reference, it is shown stationary, with the red phasor revolving around it. This situation represents a 50Ω transmission line that is connected to a purely resistive load of 150Ω. This represents a VSWR of 3:1.

- (A) The end of the line. Here, the reflected wave and the incident wave are in phase and the voltages of the two waves add to produce the resultant vector, shown in blue. Thus, the voltage of the standing wave is equal to the sum of the voltages of both the incident and the forward waves. Because the voltage of the reflected wave is half the voltage of the incident wave, the resultant at (A) will be $\times 1.5$ the voltage of the reflected wave.
- (B) 45° , or $\lambda/8$, down the line from the load, the two phasors are 90° out-of-phase, resulting in a standing wave voltage that is $\times 1.12$ the voltage of the incident wave.
- (C) 90° , or $\lambda/4$, down the line from the load, the two phasors are 180° out-of-phase, resulting in a standing wave voltage that is half the voltage of the incident wave.
- (D) 135° , or $3\lambda/8$, down the line from the load, the two phasors are again 90° out-of-phase, resulting in a standing wave voltage that is equal to $\times 1.12$ the voltage of the incident wave. (E) 180° , or $\lambda/2$, down the line from the load, the two phasors are again in phase, resulting in a standing wave voltage that is equal to $\times 1.5$ the voltage of the incident wave.

This series of phasors demonstrates the construction of the standing wave voltage from maximum through minimum and back to maximum. The cycle is repeated every $\lambda/2$ down the line.

Load mismatch

For this discussion, assume a lossless transmission line terminated with a purely resistive load of 150Ω. Let's consider the voltage levels of the two waves. For this discussion, let's assign 1.0V as the voltage of the incident wave. The voltage of the incident wave at the load will be 1.0V, but what about the voltage of the reflected wave at that point? We know that the VSWR will be $150 \div 50 = 3:1$. By using a unit called the *reflection coefficient*, we can determine the voltage of the reflected wave. Reflection coefficient can be determined from the VSWR by the formula:

$$\Gamma = \frac{S-1}{S+1}$$

where Γ = reflection coefficient
 S = VSWR.

From this we can determine that the reflection coefficient is 0.5. If we multiply the voltage of the incident wave (1.0V) by the reflection coefficient, we find that the voltage of the reflected wave is 0.5V. When the load impedance

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Table 1. The reflection coefficient, phase angle and resultant voltage of the standing wave for a purely resistive load of 150Ω in a 50Ω system.

Distance from load in wavelengths (λ)	Reflection coefficient	Phase angle	Voltage of standing wave
0λ	0.5	0°	1.5V
0.025λ	0.5	-18°	1.48V
0.05λ	0.5	-36°	1.43V
0.075λ	0.5	-54°	1.36V
0.10λ	0.5	-72°	1.25V
0.125λ	0.5	-90°	1.12V
0.150λ	0.5	-108°	0.97V
0.175λ	0.5	-126°	0.81V
0.20λ	0.5	-144°	0.66V
0.225λ	0.5	-162°	0.55V
0.25λ	0.5	-180°	0.50V
0.275λ	0.5	-198°	0.55V
0.30λ	0.5	-216°	0.66V
0.325λ	0.5	-234°	0.81V
0.35λ	0.5	-252°	0.97V
0.375λ	0.5	-270°	1.12V
0.40λ	0.5	-288°	1.25V
0.425λ	0.5	-306°	1.36V
0.45λ	0.5	-324°	1.43V
0.475λ	0.5	-342°	1.48V
0.50λ	0.5	-360°	1.50V

Table 2. The reflection coefficient, phase angle and resultant voltage of the standing wave for a purely resistive load of 16.67Ω in a 50Ω system.

Distance from load in wavelengths (λ)	Reflection coefficient	Phase angle	Voltage of standing wave
0λ	0.5	180°	0.5V
0.025λ	0.5	162°	0.55V
0.05λ	0.5	144°	0.66V
0.075λ	0.5	126°	0.81V
0.10λ	0.5	108°	0.97V
0.125λ	0.5	90°	1.12V
0.150λ	0.5	72°	1.25V
0.175λ	0.5	54°	1.36V
0.20λ	0.5	36°	1.43V
0.225λ	0.5	18°	1.48V
0.25λ	0.5	0°	1.50V
0.275λ	0.5	-18°	1.48V
0.30λ	0.5	-36°	1.43V
0.325λ	0.5	-54°	1.36V
0.35λ	0.5	-72°	1.25V
0.375λ	0.5	-90°	1.12V
0.40λ	0.5	-108°	0.97V
0.425λ	0.5	-126°	0.81V
0.45λ	0.5	-144°	0.66V
0.475λ	0.5	-162°	0.55V
0.50λ	0.5	-180°	0.50V

is greater than the system impedance, the voltage of the reflected wave is in phase with the voltage of the incident wave at the load. It is the combination (vector sum) of the incident voltage and the reflected voltage that produces the standing wave voltage. Refer to Figure 2 on page 20. Because, in this case, the reflected and forward voltages are in phase at the load, the voltage of the standing wave at this point will be 1V + 0.5V = 1.5V.

Reflection coefficient also has an

angle associated with it. This is the phase angle of the reflected wave as referenced to the incident wave. When the reflected wave is in phase, or 180° out of phase with the incident wave, the resultant standing wave voltage is determined simply by either adding the two (for in phase) or subtracting the reflected voltage from the incident voltage (for 180° out of phase).

When the phase angle is other than 0° or 180°, we must add the two, using vectors to find the resultant voltage in

the standing wave.

Reflection coefficient can also be determined easily from reflected- and forward-power measurements. Divide the reflected power by the forward power and find the square root of the result. For example, 100W forward power and 25W reflected power yields a ratio of 0.25. The square root of 0.25 is 0.5—the reflection coefficient.

Table 1 on page 22 shows the reflection coefficient, phase angle and resultant voltage of the standing wave for a purely resistive load of 150Ω in a 50Ω system. Figure 3 at the right graphs the standing wave using data in Table 1.

In the previous example, the load impedance was purely resistive and, at 150Ω, three times greater than the system impedance. Let's examine a situation where the load impedance is less than (1/3) the system impedance, or 16.67Ω. Again, the VSWR will be 3:1, and the reflection coefficient will be 0.5. Table 2 on page 22 shows the data for reflection angle, voltage of standing wave (assuming a voltage level of 1V for the incident wave) and reflection coefficient at various distances

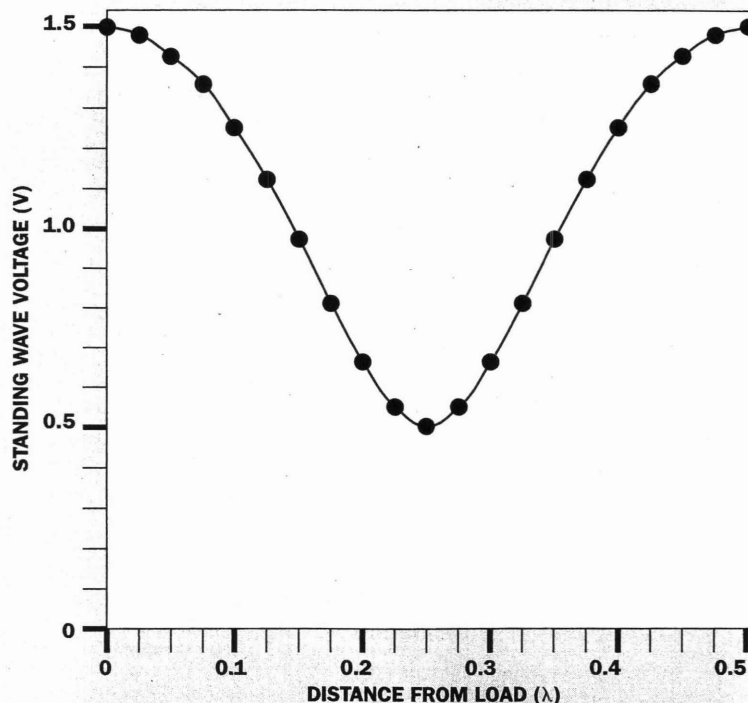
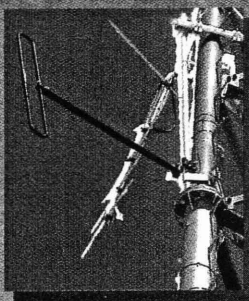


Figure 3. The impedance mismatch caused by the 150Ω load creates the VSWR pattern shown here. At the load, the voltage in the standing wave is at a maximum point (maxima). At $\lambda/4$ down the line, the voltage in the standing wave is at a minimum (minima). At $\lambda/2$ down the line, the voltage is again at a maximum. Because the VSWR is equal to the voltage maxima divided by the voltage minima, the VSWR is $1.5 \div 0.5 = 3:1$.

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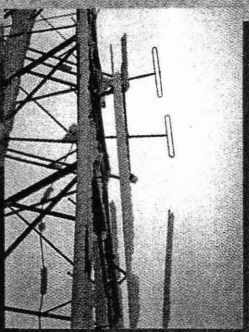
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from the load. Figure 4 at the right graphs the data from Table 2.

Compare Figures 3 and 4. When the load impedance is lower than the characteristic impedance of the transmission line, the voltages of the reflected and incident waves are 180° out of phase at the load. This results in a voltage *minima* on the standing wave pattern. The voltage maxima will occur at a point a quarter wavelength ($\lambda/4$) down the line and every $\lambda/2$ from that point. This is the reverse of the condition where the load resistance is higher than the characteristic impedance of the line. In that case, the voltage of the standing wave is at a maxima at the load because the reflected and incident waves are in phase, producing a higher resultant voltage on the standing wave pattern.

The Smith chart has long been recognized as one of the most valuable tools invented for transmission line calculations. Next month we will examine a software version of the famous Smith chart. You might be surprised to learn that the Smith chart is not for RF design engineers only.

So, until next time—*stay tuned!* ■

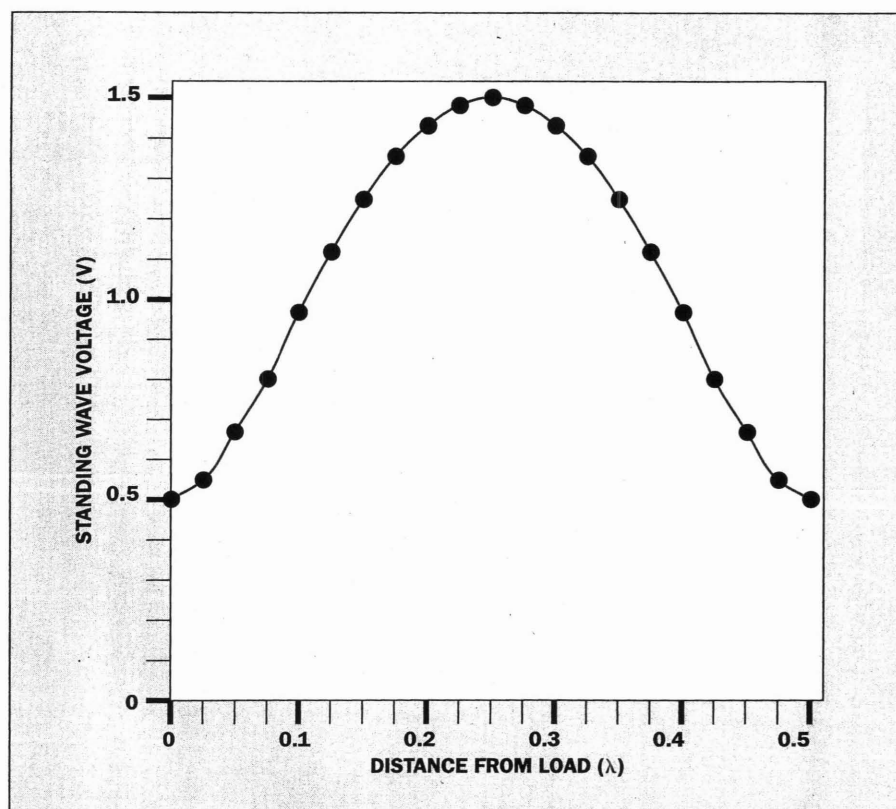
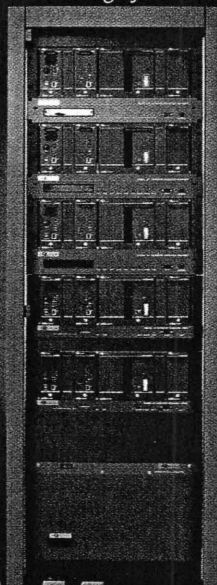


Figure 4. The impedance mismatch caused by the 16.67Ω load creates the VSWR pattern shown here. At the load, the voltage in the standing wave is at a minima. At $\lambda/4$ down the line, the voltage is at a maxima. At $\lambda/2$ down the line, the voltage is at a minima again. Again, the VSWR is equal to the maxima divided by the minima, or $1.5 \div 0.5 = 3:1$.

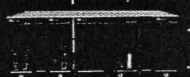
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RF amplifiers—coming and going

When precautions are followed, both receivers and transmitters can get a boost

Radio frequency amplifiers are used on both receivers and transmitters. Generally, receivers incorporate an RF amplifier stage internal to the receiver. Sometimes, an additional amplifier is placed at a point ahead of the receiver to overcome a loss such as coaxial cable loss or multi-coupler loss. Amplifiers placed ahead of the actual receiver are generally called preamplifiers.

The final output stage(s) of an RF transmitter is an RF amplifier. Frequently, an additional RF amplifier is placed between the transmitter output and the antenna to boost the RF power level. Such an RF amplifier is called a power amplifier.

In this article we will look at how both of these types of RF amplifiers are used and how they affect the communication system.

Figure 1 shows a receiver that is connected to the antenna through a long run of coaxial cable. In this example the cable has a loss of 4 dB. The noise figure of the receiver is 9 dB. The noise figure of the receiving system is equal to the cable loss plus the noise figure of the receiver (4 dB + 9 dB = 13 dB). In an ideal receiving system with a noise figure of 0 dB, the noise power in a 1 Hz bandwidth is -174 dBm. Stated another way, this is -174 dBm per hertz (-174 dBm/Hz).

This example uses a conventional FM receiver operating in the narrowband (± 5 kHz deviation) system with a sensitivity of 0.25 μ V at 12 dB SINAD. Typically, 12 dB SINAD is achieved at a signal-to-noise ratio (S/N) of approximately 4 dB. Because 0.25 μ V is -119 dBm (in a 50-ohm system), the noise floor of the receiver must be -123 dBm, approximately. Because the figure of -174 dBm/Hz is for a 1 Hz bandwidth, a correction factor must be used to adjust for other bandwidths. The correction factor is shown in Equation 1.

This receiver has a bandwidth of 15 kHz, so the correction factor is calculated using Equation 1 and found to be 41.76 dB. Thus, the noise floor of an ideal receiver with a bandwidth of 15 kHz is -132.24 dBm (-174 dBm + 41.76 dB). Because the receiver in this example has a noise floor of -123 dBm, it has a noise figure of approximately 9 dB.

With the 4 dB loss in the coaxial cable ahead of the receiver input, the system noise figure becomes 13 dB for a noise floor of -119 dBm. This means the signal at the input to the coaxial cable at the antenna has to be approximately 4 dB better, or -115 dBm.

In Figure 2 we see what can happen when we place a preamplifier ahead of the coaxial cable (for example, at the top of the tower). Suppose that the preamplifier has a gain of 15 dB and a noise figure of 3 dB. The new system noise figure becomes 4.1 dB. This is equivalent to a noise floor of approximately -128 dBm. This means that the signal input to the system must be -124 dBm to achieve 12 dB SINAD sensitivity.

Notice that this is 9 dB better than it was without the preamplifier. This is also equal to the improvement in the overall system noise figure.

Equation 1:

$$N_c = 10 \log B$$

(Where N_c is noise correction factor in decibels and B is bandwidth in hertz.)

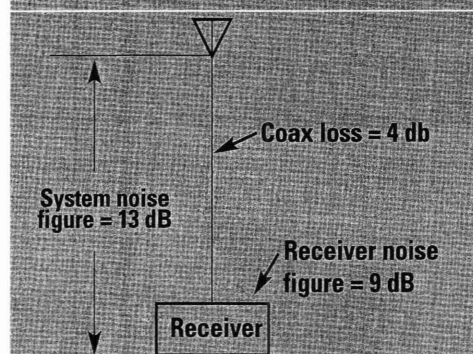


Figure 1:

The receiver with a 9 dB noise figure preceded by a coaxial cable with a loss of 4 dB yields a total system noise figure of 13 dB.

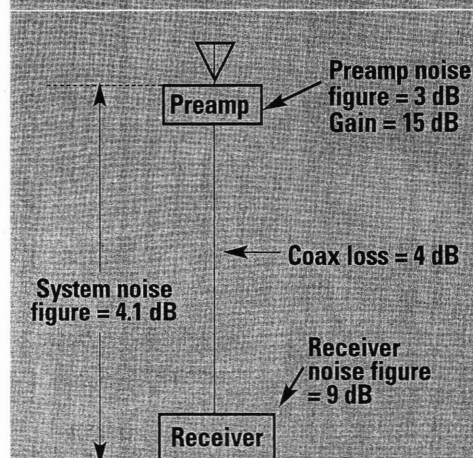


Figure 2:

With the low-noise preamplifier installed at the antenna the receiver system noise figure becomes 4.1 dB. This represents a 9 dB improvement in the overall receiver system noise figure.

A tower-top preamplifier doesn't always yield such great results. For example, if the ambient noise level at the communication site is high, the improvement generated by a tower-top preamplifier is less dramatic. Generally, the higher the ambient site noise level, the smaller the net improvement offered by the preamplifier. With high site noise levels, a preamplifier is practically useless, as the ambient noise is amplified alongside the signal, resulting in no improvement in the signal-to-noise ratio. The signal-to-noise ratio at the input to the receiving system is the *real* measure of improvement.

Try to choose an RF preamplifier with a low noise fig-

The higher the ambient site noise level, the smaller the improvement offered by the preamplifier. With high site noise levels, a preamplifier is practically useless.

ure and a high third-order intercept point (this is a measure of the intermodulation rejection figure of the amplifier). In high signal-level areas, preamplifier use may result in more severe intermodulation and receiver-desense problems. Intermodulation products formed in the receiver may be just below the threshold of interference without the preamplifier. With the preamplifier in line, the levels of the input signals will drive the intermodulation product(s) to a level sufficient to cause nuisance or destructive interference. The same thing is true of desense. Potential desense problems may lie just beneath the surface without the preamplifier but become quite apparent after installation. Be aware of these potential problems before deciding whether to use a preamplifier.

A power amplifier also may be used to boost the RF output power level of the transmitter. For example, the output of a handheld unit installed in a vehicular adaptor is only a few watts. However, by using an RF power amplifier, the output level can be boosted to a level comparable to a regular mobile transceiver. Or, a 10- to 25-watt base station can be boosted to a much higher power level using an RF power amplifier.

RF power amplifiers can be classified as broadband or narrowband. The broadband units usually need no tuning when used at the proper frequency. However, narrowband units will need to be tuned for the particular frequency or

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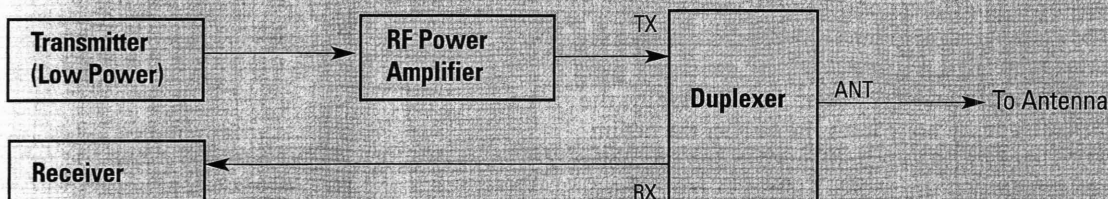
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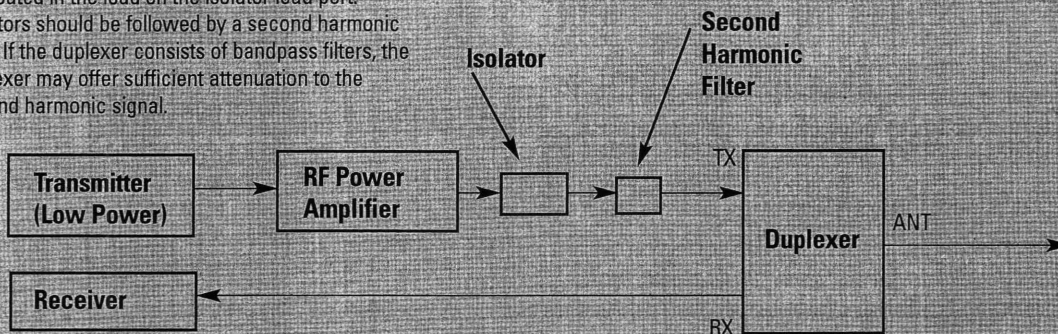


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This is a typical repeater arrangement. The RF amplifier may not like the load presented by the input to the duplexer. The line length between the duplexer and the RF amplifier output may be changed to improve the load impedance to the RF power amplifier.

An isolator is used to present a proper 50-ohm impedance to the RF power amplifier. Although a mismatch may still exist, the RF power amplifier sees a 50-ohm impedance. Any reflected power will be dissipated in the load on the isolator load port. Isolators should be followed by a second harmonic filter. If the duplexer consists of bandpass filters, the duplexer may offer sufficient attenuation to the second harmonic signal.



narrow band of frequencies over which the amplifier must operate. In the case of a transceiver, the RF amplifier must have a bypass that allows the receive signal to pass through the amplifier in the reverse direction with little attenuation.

RF power amplifiers are designed to work into a purely resistive load of 50 ohms. Highly reactive loads—such as an improperly tuned antenna, cavity filters and duplexers—may cause the amplifier to become unstable. Sometimes the length of the coaxial line between the RF amplifier and the duplexer or antenna can be changed to reflect a better load to the RF amplifier output. Changing the line length simply tunes out the reactive component of the load impedance.

Figure 3 shows a typical repeater/base station using an external RF power amplifier. Changing the line length between the duplexer and RF power amplifier sometimes helps to

present a better load condition to the output of the amplifier.

An alternative is shown in Figure 4.

Using precautions associated with receiver preamplifiers and transmitter power amplifiers can prevent many headaches.

An RF isolator can be used between the RF amplifier output and the duplexer or load. The output of the RF amplifier “sees” a good 50-ohm match even though a mismatch may exist between the isolator and the load. Any reflected

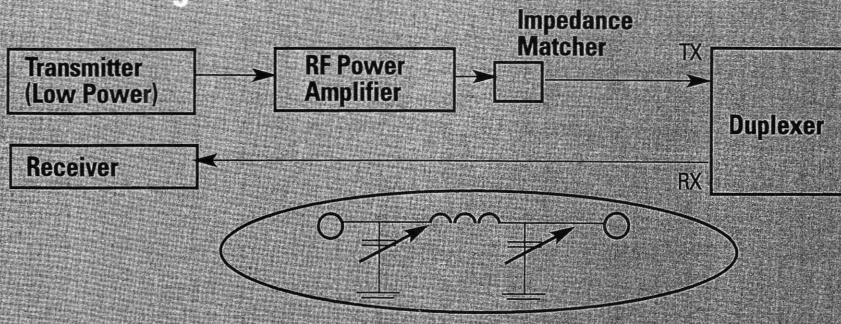
power is dissipated in the load resistor connected to the load port of the isolator. Normally, a second harmonic filter follows the output of the isolator, since ferrite materials tend to generate second harmonics. Don’t connect an isolator to the output of a transceiver—it will block the receiver input. The isolator must be connected only to the transmitter or power amplifier output.

Yet another alternative is shown in Figure 5. Here, an impedance matcher (pi-network) is connected between the output of the RF amplifier and the load impedance. The pi-network can be tuned so that the RF amplifier “sees” a 50-ohm load impedance.

In addition to understanding how RF amplifiers may be used to improve the performance of a communication system, it is also crucial to understand how to properly use them. Important precautions that should be followed include:

- Do not overdrive the input

Figure 5:



An impedance matcher such as the pi-network shown in the inset can be used to provide a proper match to the RF power amplifier. Such impedance matchers are designed to mount right on the output connector of the RF amplifier. Tuning adjustments are accessible through small holes usually covered by caps to keep dust and other foreign matter out of the device.

of the amplifier

- Make sure the amplifier "sees" a proper 50-ohm load impedance
- Do not exceed the DC input current or voltage rating of the amplifier
- Make sure the amplifier has adequate ventilation in mobile installations
- Make sure the amplifier cooling fans are operating
- Don't just check the power output with a wattmeter; check it with a spectrum analyzer

Using the proper precautions associated with both receiver preamplifiers and transmitter power amplifiers can prevent many headaches. In the proper situation, both of these devices can improve the communication system. Used improperly, either of these devices can cause serious problems.

Until next time—stay tuned! ■

Harold Kinley's latest book, "*The Radioman's Manual of RF Devices*," is available from Noble Publishing. Visit www.noblepub.com to order.

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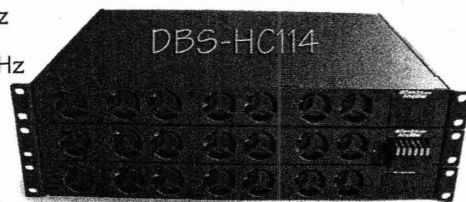
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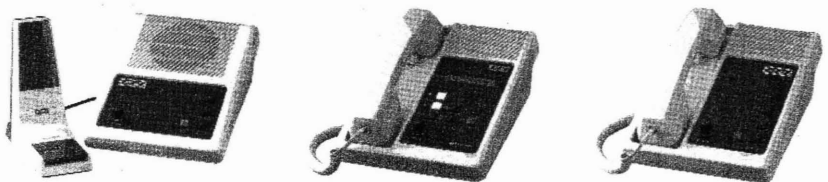
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Fine-tuning for the RF isolator

The goal is minimum insertion loss in the forward direction, maximum isolation in the reverse

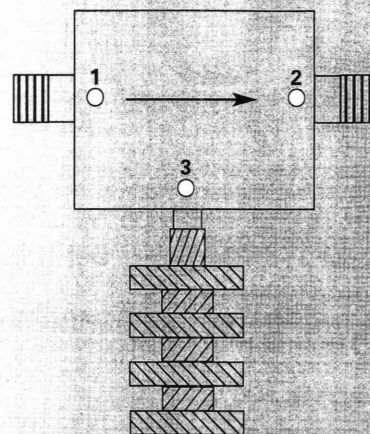
Last month, the operation of RF isolators was discussed in detail. This article focuses on the proper tuning of the RF isolator. Several techniques can be used, depending upon the type of test equipment available. Preferably, the isolator should be tuned in its final installed position. The objective is to tune the isolator so it presents the minimum insertion loss in the forward direction and maximum insertion loss (isolation) in the reverse direction.

Figure 1 shows a block diagram of a simple single isolator. The arrow indicates forward flow of RF power. Any reverse or reflected power flow will be dumped into the RF dummy load at port 3. Any reflected power at port 3 will be transferred back to port 1, the input port. It is important that the trimmer at the input port (port 1) is tuned for minimum reflected power at port 1 so that the transmitter "sees" very little reflected power. There will always be some residual reflected power at the input port, even when the load connected to the output port (port 2) is a perfect 50-ohm load.

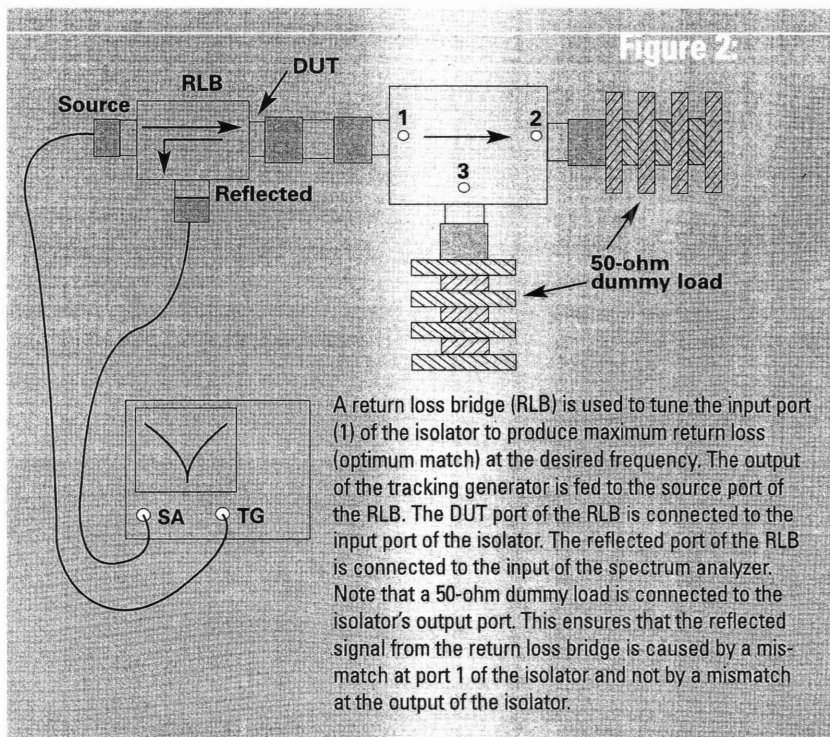
One method of tuning the input port is to use a return loss bridge (RLB). The test setup for using this method to tune the input port is shown in Figure 2. It is important to terminate the output of the isolator with a 50-ohm dummy load. This ensures that the reflected signal at the reflected port of the RLB is caused predominately by the mismatch at port 1 of the isolator and not by any reflected signal from the output port of the isolator.

A simple method of tuning a single isolator is shown in Figure 3. A through-line wattmeter is connected between the output port of the isolator

and a 50-ohm dummy load. Using the setup in Figure 3A, the input and output ports (1 and 2) are tuned for maximum power on the wattmeter. Using the setup in 3B, the load port (3) is tuned for minimum power on the wattmeter. Using the setup at B, it is important that the transmitter power not exceed the power rating of the dummy load connected to port 3 of the isolator. Also, when using the setup in 3B, it may be necessary to use a low-power element in the wattmeter to tune for minimum power. Actually, when using this method, tuning the isolator to the deepest null may be difficult, at best. Suppose that the isolator has an isolation of 30 decibels. If the transmitter output power is 100 watts, then the power appearing at port 2 will only be 0.1 watts or 100 milliwatts. Obviously, the wattmeter element would have to



Simple drawing of a single RF isolator. Forward flow of RF power is from port 1 to port 2.



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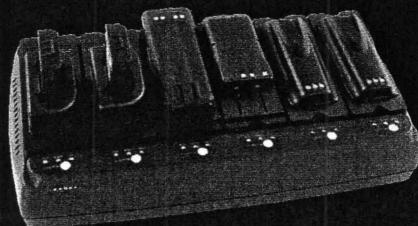
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be very sensitive in order to get a sufficient reading to properly tune port 3 for the deepest null.

A better way to tune port 3 for maximum isolation is shown in Figure 4. The logarithmic display of the spectrum analyzer/tracking generator allows much greater range for tuning the isolator for maximum isolation.

The situation gets more interesting when a dual isolator is tuned (Figure 5).

trimmer at port 6 is adjusted for maximum isolation (deepest null). Then, load 2 is removed and the trimmer at port 5 is adjusted for maximum isolation. Never adjust the trimmer at the port where the load is removed. Rather, always adjust the trimmer with the load connected to that port. With both loads connected, trimmers at ports 5 and 6 may be tweaked for maximum isolation.

If you don't have a tracking genera-

Figure 3A:

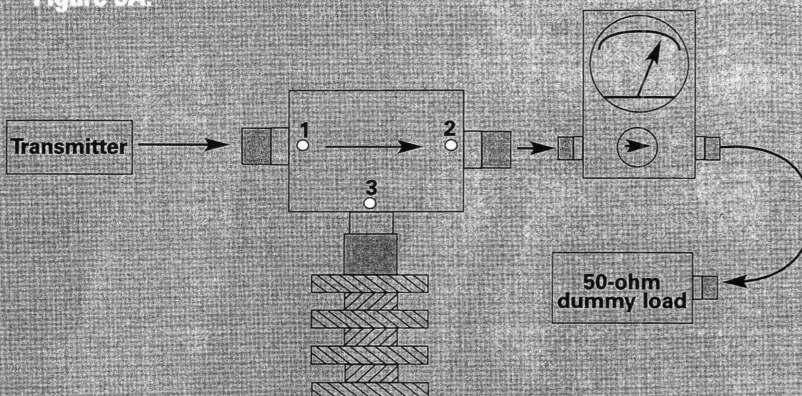
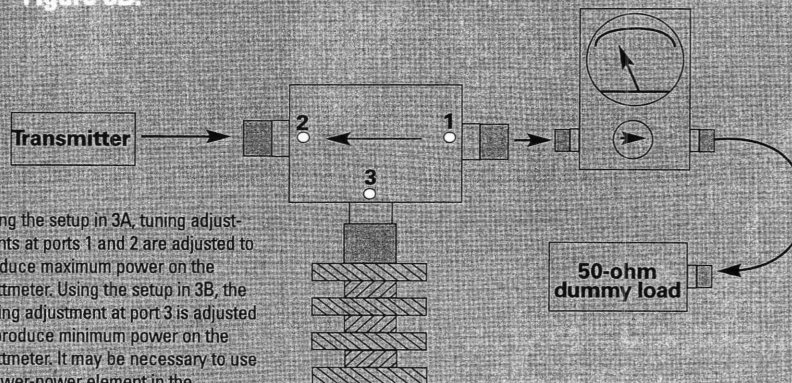


Figure 3B:



Using the setup in 3A, tuning adjustments at ports 1 and 2 are adjusted to produce maximum power on the wattmeter. Using the setup in 3B, the tuning adjustment at port 3 is adjusted to produce minimum power on the wattmeter. It may be necessary to use a lower-power element in the wattmeter to get a useable indication.

The forward response of the isolator is tuned by adjusting the trimmers at ports 1, 2, 3 and 4. The peak of the response should be at the operating frequency, and the curve should be symmetrical. The forward response of the isolator is very broadband. Check the insertion loss at the operating frequency against the manufacturer's specifications.

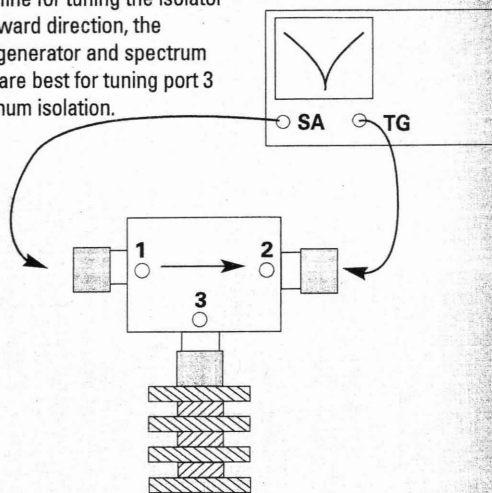
The reverse response or isolation tuning is done using the setup shown in Figure 6. First, load 1 is removed, and the

tor/spectrum analyzer combo, you can use a signal generator with a spectrum analyzer. You won't be able to get a swept response but you will be able to tune for minimum insertion loss in the forward direction and maximum isolation in the reverse direction. If you don't have a spectrum analyzer, you can use a receiver tuned to the correct operating frequency. Even a scanner can be used.

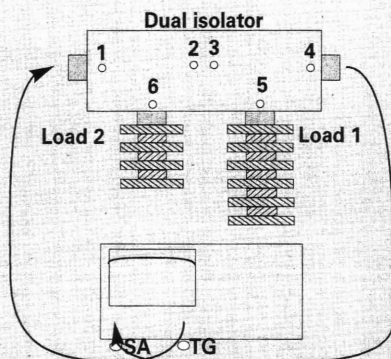
To tune for minimum insertion loss in the forward direction, an AC voltmeter

TECHNICALLY SPEAKING

While the through-line wattmeter might be fine for tuning the isolator in the forward direction, the tracking generator and spectrum analyzer are best for tuning port 3 for maximum isolation.



The trimmers at ports 1, 2, 3 and 4 are tuned for maximum on the spectrum analyzer. The response curve should be symmetrical with the maximum response at the operating frequency. The insertion loss should be compared with the manufacturer's specifications.

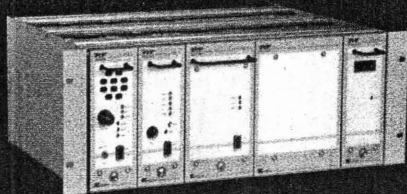
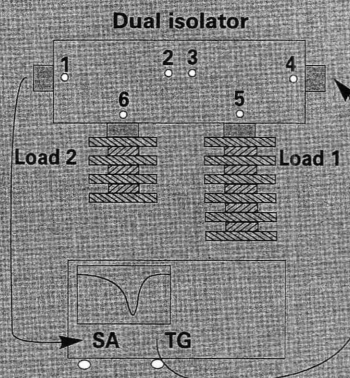


is connected across the speaker terminals of the FM receiver (Figure 7). Then, the signal generator is set to produce a measurable amount of noise across the speaker terminals. All the input/output ports are tuned for minimum noise level on the voltmeter. As these ports are tuned, it may be necessary to reduce the signal generator level to maintain a useable noise level at the receiver output. To get a reference benchmark, the signal generator is connected directly to the FM receiver and adjusted to produce a reference noise level on the AC voltmeter. The reference noise level as well as the signal generator level should be recorded. The insertion loss of the isolator can be determined by comparing the signal generator level required to produce the reference noise level with the recorded benchmark.

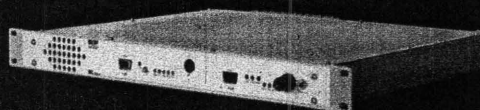
To tune for maximum isolation, the connections to the isolator are reversed. Then, the signal generator is set to produce a measurable noise level on the AC voltmeter. The load ports are then tuned, as previously described, for maximum noise level on the AC voltmeter. As the load ports are tuned, it may be necessary to increase the signal generator

Figure 6:

This setup is used to tune the isolator for maximum isolation. First, load 1 is removed and port 6 tuned. Then, load 2 is removed and port 5 tuned. See text for details.



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level to reduce the noise level at the receiver output. The isolation can be determined by comparing the signal generator level required to produce the reference noise level with the recorded benchmark.

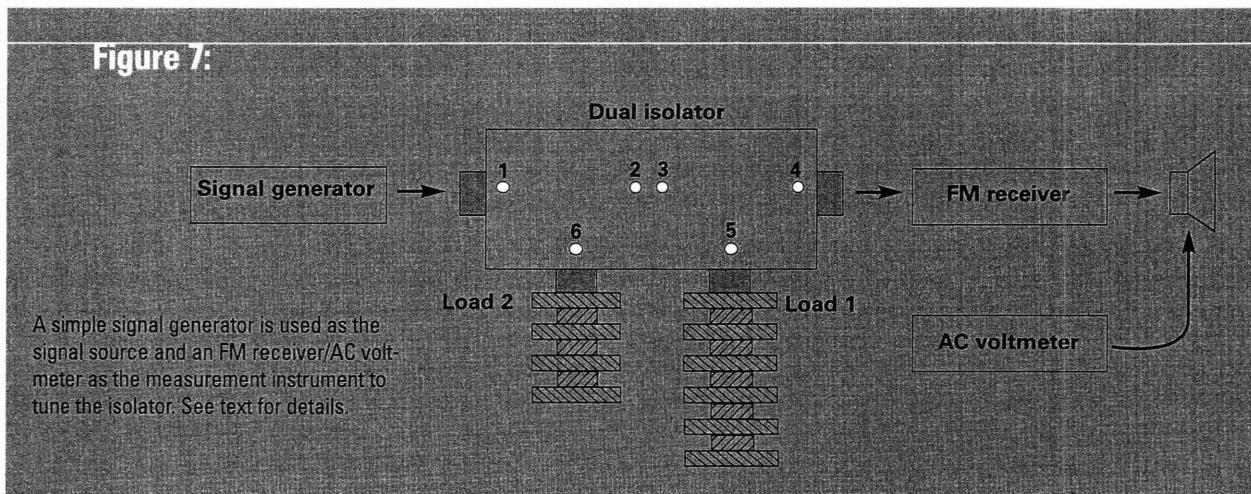
The exact method you use to tune

the isolator will be determined by the type of test equipment you have available. It is best to tune the isolator in the final installed position. Some isolators might be detuned by installing them on a ferrous metal surface. Check the manufacturer's instructions to avoid

improper installation. Minimum insertion loss in the forward direction and maximum isolation in the reverse direction is the goal. An occasional check of these performance specifications is recommended.

Until next time—stay tuned! ■

Figure 7:



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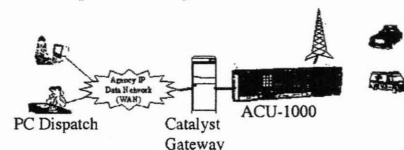
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The isolator: An RF traffic director

It effectively blocks the return of reflected power when basic precautions are taken

An RF isolator keeps reflected power from returning to the transmitter output and keeps other signals from getting into the transmitter output. It also provides a good impedance match to the transmitter output. This article describes the basic RF isolator and how it is used.

Figure 1 shows a drawing of a typical RF isolator. It is a

three-port device with an input port, an output port and a load port. For our purposes, there is little need to know the internal design of the RF isolator. Hence, the RF isolator will be treated as a black box in this discussion. As depicted by the solid arrow in Figure 1, the normal direction of RF flow is into the input port, through the isolator, and out of the output port. Any RF signal entering the output port is directed through the isolator to the load port where it is dissipated as heat in the dummy load. The dashed arrow in Figure 1 depicts the reverse RF flow.

Figure 2 shows the typical response of the RF isolator in the forward and reverse direction. The response in the forward direction is very broad, offering little attenuation to a broad range of input frequencies. However, in the reverse direction, the response is relatively narrow. Maximum attenuation occurs at a single frequency or a very narrow band of frequencies.

Using the black-box approach, let's examine the typical operation of a properly tuned isolator (Figure 3). An RF isolator is connected to (and physically located near) the output of the transmitter. The green arrow depicts the normal forward RF flow. The dashed red arrow depicts the reverse RF flow back into the isolator. This reverse flow could be reflected power from the antenna, extraneous signals picked up by the antenna, or both. The RF signal entering the output port of the isolator is dumped off into the dummy load at the load port. The dummy load then dissipates the returned power as heat.

Let's analyze the operation of an RF isolator by applying specific numbers to the setup shown in Figure 3. In this example, the transmitter RF output power is 100 W, the insertion loss of the isolator (in the forward direction) is

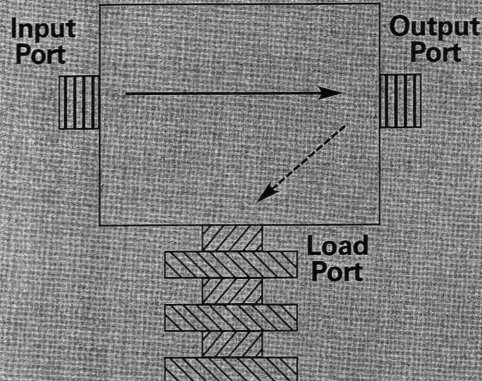


Figure 1:

A typical single-section isolator is shown here. Normal RF flow is from the input port on the left to the output port on the right. Reflected power, indicated by the dashed arrow, is 'dumped off' into the dummy load connected to the load port.

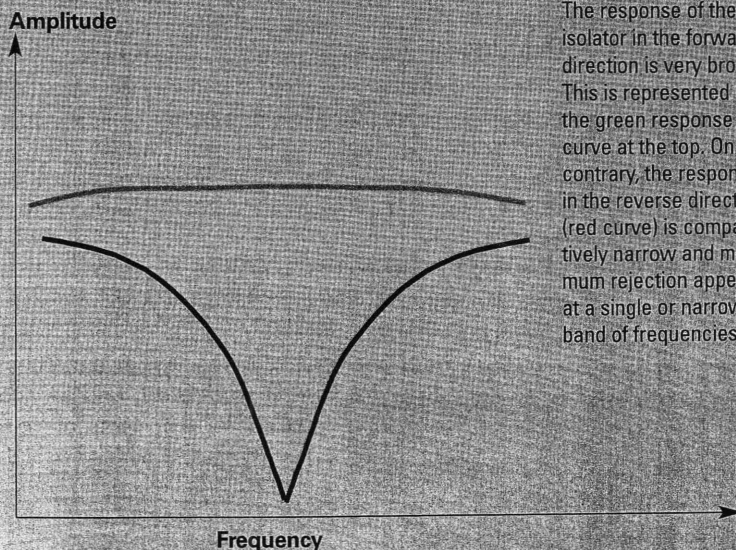


Figure 2:

The response of the isolator in the forward direction is very broad. This is represented by the green response curve at the top. On the contrary, the response in the reverse direction (red curve) is comparatively narrow and maximum rejection appears at a single or narrow band of frequencies.

0.3 dB, and the transmission line loss is 1.5 dB at the operating frequency. Let's also establish that the SWR at the antenna is 3:1 while the isolation (reverse flow insertion loss) of the RF isolator is 30 dB. Because the isolator has an insertion loss of 0.3 dB in the forward direction, the output power is 93.3% of the input power to the isolator. Consequently, we are able to demonstrate in Equation 1 (on page 36) that the forward power at the output of the isolator is 93.3 W.

Table 1 shows the equivalent percentage of power for decibel losses from 0.1 to 1 dB. The forward power at the antenna can be calculated using Equation 1. Or the calculation can be done using the data from Table 1. Because the transmission line has a loss of 1.5 dB, we can use the multiplication factors of 1 dB and 0.5 dB in Table 1 to calculate the forward power at the antenna. In this case, the factor for 1 dB is 0.794 and for 0.5 dB is 0.891. Multiplying these two factors and then multiplying the result by the forward power at the isolator output yields forward power at the antenna of 66 W [$0.794 \times 0.891 \times 93.3$].

The next step is to calculate the reflected power from the antenna, which we do in Equation 2 (on page 36). Hence, the reflected power is 25% of the forward power, or 16.5 W [0.25×66 W]. The reflected power will incur the same 1.5 dB of line loss. Using Table 1 as before, we find the reflected power at the output port of the isolator to be 11.7 W [$0.794 \times 0.891 \times 16.5$ W]. If the internal loss between the output port and the load port is negligible, then the power applied to the dummy load will be approximately the same as the reflected power at the output port, or 11.7 W.

Because the isolation (reverse insertion loss) of the RF isolator is specified as 30 dB at the operating frequency, the reflected power appearing at the RF output of the transmitter will be 11.7 W minus 30 dB. Because a reduction of

10 dB represents a power multiplication factor of 0.1, then 30dB is a multiplication factor of 0.001. So the reflected power appearing at the transmitter output equals 11.7 mW [0.001×11.7 W = 0.0117 W]. Table 2 offers a handy chart for finding the equivalent power reduction for decibel

It is important to note that even under perfect load conditions at the output port of the isolator there will be some reflected power at the input port of the isolator due to a slight mismatch at the port.

losses. The net radiated power is the forward power at the antenna minus the reflected power at the antenna, or 49.5 W [66 W - 16.5 W].

It is important to note that even under perfect load conditions at the output port of the isolator there will be some reflected power at the input port of the isolator due to a slight mismatch at the port. In the case of the preceding example, suppose that the manufacturer's specifications

Figure 3:

An RF isolator is connected to the output of the transmitter. The green arrow depicts the direction of normal forward RF flow. Any reflected power, depicted by the red dashed arrow, is returned to the isolator and "dumped" into the dummy load connected to the load port.

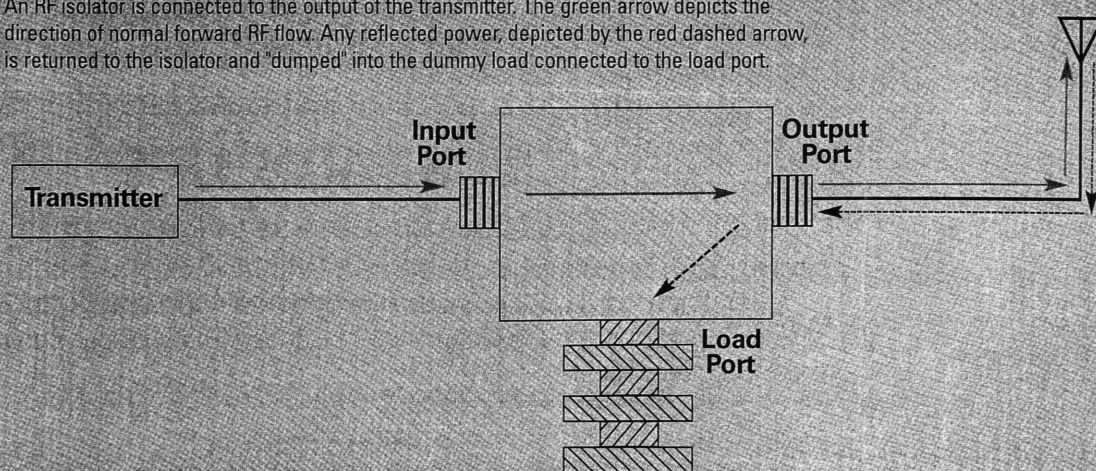
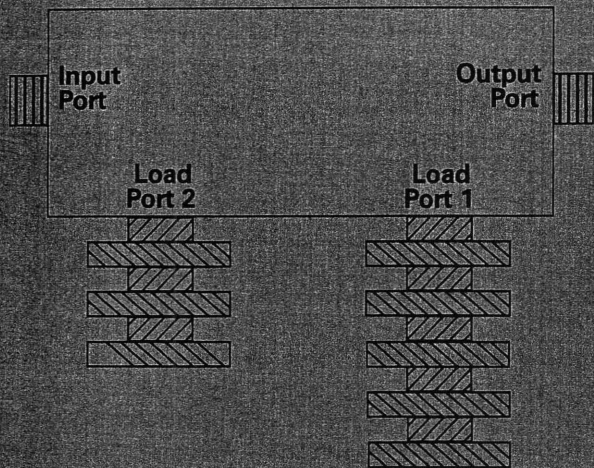


Figure 4:

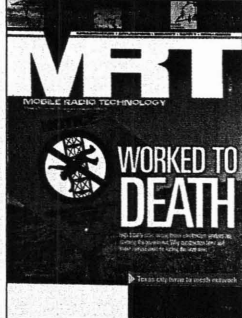
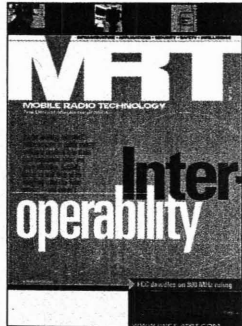
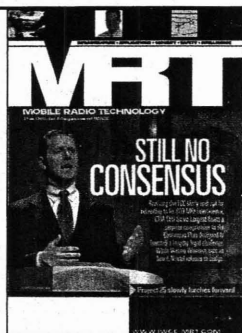
A dual-section RF isolator is depicted here. The dual-section isolator has approximately twice the isolation of a comparable single isolator and twice the insertion loss in the forward direction. Notice that the dummy load connected to the load port nearest the output port is larger than the other dummy load. This is because this is the first load port that the reflected power sees and therefore is the first line of defense. Any power reflected from load port 1 is "seen" at load port 2. Generally, the dummy load at the first load port is rated at half the transmitter power. For a 100 W transmitter, the rating of the dummy load would typically be 50 W.



stated that the SWR at the input port is 1.15:1 or better. Assuming a worst-case of 1.15:1, the reflected power under perfect load conditions at the isolator output would be 0.4867 W or 486.7 mW.

The combination of both reflected powers, the reflected component caused by the mismatch at the antenna and the reflected component caused by the mismatch at the isolator input port, amounts to a difference of only 0.1 dB, approximately. Thus, the most significant contribution of reflected power, in this example, is caused by the isolator input port mismatch.

To further explore the significance of the input port mismatch, assume a worst-case mismatch at the output of the isolator, for instance a shorted connector attached to the isolator output port. As before, 100 W is applied to the input port. With a 0.3 dB insertion loss, 93.3 W appears at the output



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port. With an isolation of 30 dB, the reflected power at the input port (caused by the mismatch at the output port) will be 0.0933 W or 93.3 mW. This is still much less than the reflected power caused by the mismatch at the input port and will increase the total reflected power seen at the transmitter by less than 1 dB. Note: the dummy load at the load port usually is not large enough to handle this amount of reflected power.

Of course, the above discussion is theoretical. In the real world things change a bit. First, there is the heat factor. The heating caused by normal operation can compromise the performance of the RF isolator. Excessive reflected power will increase the required load dissipation and increase the temperature of the isolator. Consequently, reflected power should be kept to a minimum. Even though the isolator prevents the transmitter from "seeing" most of

the reflected power, the isolator is not an impedance matching device. This is worth repeating: the RF isolator is not an impedance matching device! Even so, the RF isolator provides an excellent 50-ohm load impedance to a transmitter.

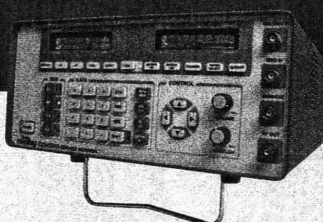
Manufacturer's specifications and recommendations should be followed in order to ensure proper operation of the isolator. If the manufacturer states that the isolator should not be mounted on a ferrous metal surface, doing so might significantly detune the isolator and reduce the performance. Also, do not exceed the power input specification rating of the isolator. Power input should be reduced commensurately to keep reflected power at the load port within the limits of the dummy load.

Dual- and even triple-section isolators are available for even greater isolation (Figure 4). Single isolators can be

Even though the isolator prevents the transmitter from "seeing" most of the reflected power, the isolator is not an impedance matching device.

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placed in cascade to achieve higher isolation than afforded by a single isolator, but this requires more space and interconnecting cables. Better heat dissipation is one advantage of this configuration. Also, should a single isolator in the cascade fail, it can be replaced at less expense than replacing a dual- or triple-section isolator. Remember that while isolation figures increase, so too does the forward insertion

loss with multiple sections or cascades.

Make sure adequate ventilation is provided, even if it means adding an external fan to cool the isolator. Sometimes, the dummy load can be placed away from the isolator via an interconnecting cable. This can reduce the heat near the isolator in the case of significant reflected power. The dummy load should be rated for at least half the trans-

mitter power. For a 100 W transmitter, the dummy load should be rated for at least 50 W. For multi-section isolators, the dummy load connected to the first load port should be rated for at least 50% of the transmitter power output. The remaining loads can be much smaller since they don't see as much reflected power.

Be sure to follow the output of the isolator(s) with a second-harmonic filter. Do

Equation 1:

$$P = 10^{\left(\frac{-\text{dB}}{10}\right)} \times 100 = 10^{\left(\frac{-0.3}{10}\right)} \times 100 = 0.933 \times 100 = 93.3\%$$

Equation 2:

$$R = \left(\frac{S-1}{S+1} \right)^2 = \left(\frac{3-1}{3+1} \right)^2 = \left(\frac{2}{4} \right)^2 = (0.5)^2 = 0.25$$

where R is reflected power and S is SWR

Table 1:

dB loss	Percent power
0.1	97.7
0.2	95.5
0.3	93.3
0.4	91.2
0.5	89.1
0.6	87.1
0.7	85.1
0.8	83.2
0.9	81.3
1.0	79.4

Table 2:

dB loss or isolation	Multiplication factor
1	0.794
2	0.631
3	0.501
10	0.1

By using common sense and following recommendations and specifications, you can get maximum performance and life out of a well-built RF isolator.

not place the isolator at the output of a transmitter combiner. Each individual transmitter should be followed with an isolator and harmonic filter. Isolators should be used as a matter of standard procedure in order to keep intermod-forming signals out of the transmitter output stage or to significantly reduce the level of intermod products from the transmitter. By using common sense and following manufacturers' recommendations and specifications, you can get maximum performance and life out of a well-built RF isolator. Next month we will look at field-tuning isolators using various techniques and equipment.

Until next time—stay tuned! ■

The finer points of rebanding

Follow these tips to eliminate receiver intermodulation and out-of-band emissions

Last month's Tech Speak focused on techniques for measuring interference in the 800 MHz band, both during and after the rebanding process, and touched upon strategies to mitigate that interference. This month, we examine mitigation techniques in greater detail.

Before the FCC last summer ordered Nextel Communications to reband the airwaves over the next three years, many licensees questioned whether rebanding by itself actually would eliminate interference. These skeptics were correct as rebanding by itself does not eliminate interference. Instead, it creates necessary conditions so that two important mitigation techniques become effective. This article describes how Nextel can put these mitigation techniques to use after rebanding.

Let's begin with a refresher on the types of interference created by Nextel and the A-Band cellular carrier. This interference affects the portable radio receiver (not the repeater receiver) and falls into two categories:

- Receiver intermodulation is a non-linear combination of two or more interfering signals inside the receiver front-end (low-noise amplifier and/or mixer).

- Out-of-band emissions (OOBE) comprise radio frequency energy that falls outside the assigned channel

for the transmitter. OOBE includes radio carrier harmonics, transmitter intermodulation products, and broadband transmitter noise typical of digital radio transmitters.

A block diagram for a typical public-safety radio receiver is shown in Figure 1. Public-safety receivers are vulnerable to receiver intermodulation because the front-end bandpass filter must pass all frequencies from 851 MHz to 869 MHz. Nextel frequencies fall within this same band, so the low-noise amplifier that follows the bandpass filter is exposed to strong interfering signals that mix within the amplifier.

An intermodulation, or IM, product between two interferers with frequencies f_1 [lower

frequency] and f_2 [higher frequency] can be represented mathematically, as depicted in Equation 1. Using this formula, two interferers operating at 861.4875 MHz and 862.4875 MHz will create the third-order IM products inside the 800 MHz band depicted in Equation 2.

An important rule-of-thumb for two-transmitter, third-order products is that the products will fall exactly $\Delta = f_1 - f_2$ below the lower frequency and above the higher frequency. It also should be noted that three interferers also can create third-order products, which take the form $A+B-C$ rather than $2A-B$, and that three-carrier products are more numerous than two-carrier products, as shown in Equation 3.

Generally—given a set of poten-

Equation 1

$$f_{im} = nf_1 + mf_2$$

Where n and m are non-zero integers. The order of the product is simply the sum of the absolute values of the coefficients, $|n| + |m|$.

Equation 2

$$2 (861.4875 \text{ MHz}) - 862.4875 \text{ MHz} \\ = 860.4875 \text{ MHz}$$

$$2 (862.4875 \text{ MHz}) - 861.4875 \text{ MHz} \\ = 863.4875 \text{ MHz}$$

Figure 1

Typical public-safety receiver front end

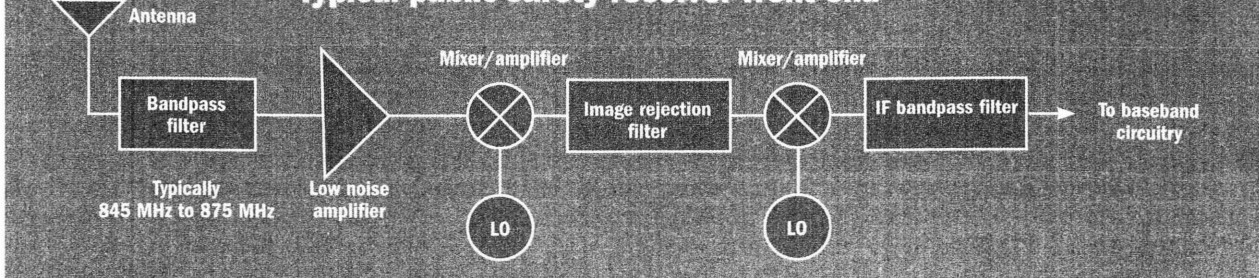
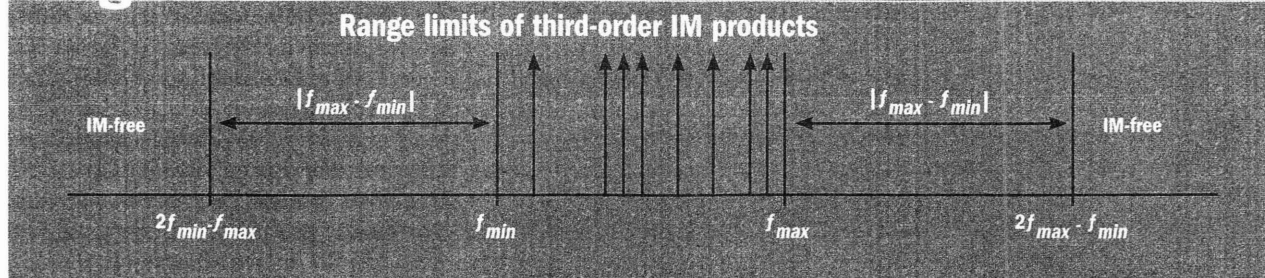


Figure 2

tial interfering frequencies that fall within the range $[f_{min}, f_{max}]$ —no third-order product can fall farther than $\Delta = f_{max} - f_{min}$ below f_{min} or above f_{max} . A sufficient condition to preclude third-order IM products is stated mathematically in Equation 4 on page 22, while this principle is shown graphically in Figure 2. Note that this article is focusing on third-order products exclusively. While fifth-order products do occur, they usually are weak enough in a Class A receiver that they can be ignored.

Given Figure 2, a simple way to eliminate receiver intermodulation is to first create a guard band between dissimilar services and then

Equation 3

$$\text{Three-carrier products} = N(N-1)(N-2)/2$$

$$\text{Two-carrier products} = N(N-1)$$

Where N is the number of transmit frequencies.

ensure frequency sets for the interfering service (e.g., Nextel) have a span no greater than the span of the guard band. Because guard bands waste spectrum, it is important to limit the number of band edges between the two dissimilar services so only one guard band is required. This is precisely what the FCC has done with its rebanding plan.

The 851 MHz to 869 MHz band now

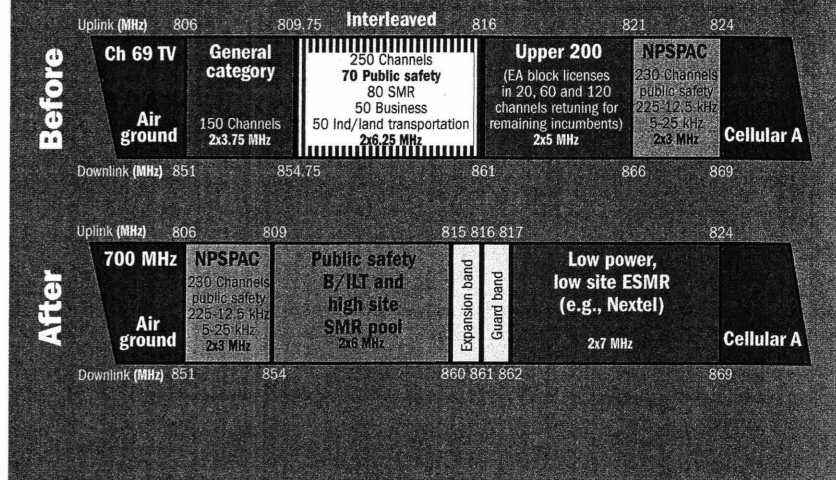
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Figure 3

FCC spectrum allocation



consists of four sub-bands as shown in the before segment of Figure 3.

Note that public-safety channels occur in the interleaved band and in the NPSPAC band. Interleaving alone creates insurmountable problems, but the four band edges also create problems. The proposed rebanding in the after segment of Figure 3 reduces the band edges from four to one and effectively creates a 2 MHz guard band where Nextel has agreed to certain significant service restrictions.

The 2 MHz between 860 MHz and 862 MHz is subdivided into an expansion band (860 MHz to 861 MHz) and a guard band (861 MHz to 862 MHz). No public-safety system is required to remain in or relocate to the expansion band, although they may do so voluntarily. The level of protection afforded in the expansion band creates the effect of a pseudo-guard band, which has a twofold purpose: It allows bandpass filters at the cell site to roll off, and it enables Nextel to create channel plans with intermodulation products that fall in the guard band, but not on public-safety channels.

After rebanding, the solution to OOB is straightforward. Each of Nextel's transmit antenna sectors in potential problem areas should have a cavity bandpass filter designed to pass 862 MHz to 869 MHz and reject

Equation 4

$$f_r < 2f_{min} - f_{max} \text{ or } f_r > 2f_{max} - f_{min}$$

Where f_r is the nearest receive frequency requiring interference protection.

Equation 5

$$2f_{min} - f_{max} > 860 \text{ MHz}$$

Set	Span	Channels	f_{min} (MHz)	f_{max} (MHz)	Lowest IM sector (MHz)	Lowest IM cell (MHz)
A1	675 kHz	27	862.0125	862.6625	861.3625	
A2	675 kHz	27	862.6875	863.3375	862.0375	860.0125
A3	675 kHz	27	863.3625	864.0125	862.7125	
B1	825 kHz	33	864.0375	864.8375	863.2375	
B2	825 kHz	33	864.8625	865.6625	864.0625	861.5875
B3	825 kHz	33	865.6875	866.4875	864.8875	
C1	825 kHz	33	866.5125	867.3125	865.7125	
C2	825 kHz	33	867.3375	868.1375	866.5375	864.0375
C3	850 kHz	34	868.1625	868.9875	867.3375	

Potential frequency sets (compatible with N=3 re-use) (A, B, C=cell; 1, 2, 3=sector)

all frequencies below 860 MHz by at least 45 dB.

The solution to receiver IM also is straightforward. The objective is to create channel sets using all of Nextel's rebanded channels (280 total) without creating any third-order IM products on or adjacent to a public-safety channel. For each Nextel channel set with a minimum frequency $[f_{min}]$ the maximum frequency $[f_{max}]$ must obey the rule depicted in Equation 5.

There are many feasible solutions to this problem. Using a seven-cell, sectored re-use pattern with 280 total channels, Nextel only requires 13 channels per sector $[280/7/3]$. A three-cell reuse pattern requires 31 channels per sector. For an iDEN system, three-cell re-use is likely to be a worst-case scenario.

We know that Nextel can combine adjacent channels on the same sector antenna using hybrid combiners. Thus, 13 channels require only 325 kHz and thirty-one channels require 775 kHz. But the proposed effective guard band is 2 MHz, so feasible solutions exist. A potential frequency set is listed in Table 1 on page 23.

Note from Table 1 that we have limited the range of frequencies in each cell site (three sectors total) so IM products that would occur between sectors still fall above 860 MHz. Thus, a public-safety receiver operating in the overlap region between sectors is still protected.

Channel sets A1, A2 and A3 have less than 31 channels, but these sets can be assigned to relatively low-capacity sites.

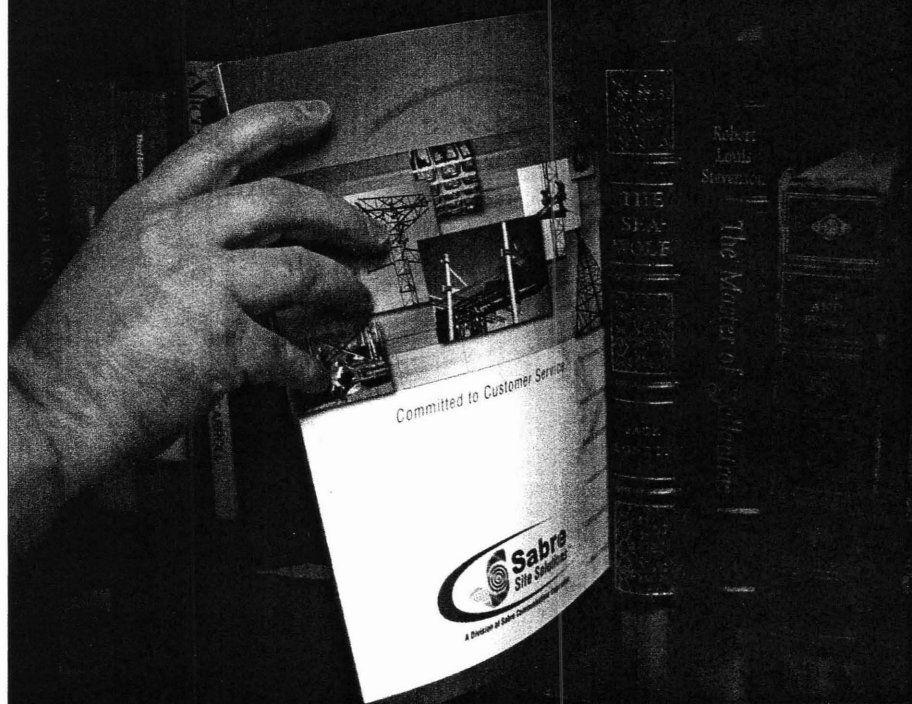
Table 1 is only one feasible solution. There are literally hundreds of practical solutions that meet Nextel's capacity needs while fully protecting public safety from all third-order IM products. Any channel set that obeys Equation 5 within a sector and between adjacent sectors is a feasible solution. Of course, Nextel does not consult the author when preparing its

frequency plan, so the company might take a different approach.

So, is interference eliminated entirely? No. There are still isolated opportunities for interference to occur, especially at Nextel/A-Band co-location sites. These require special attention and exist to keep the consulting engineer gainfully employed. ■

Jay Jacobsmeyer is president of Pericle Communications Co., a consulting engineering firm located in Colorado Springs, Colo. He holds bachelor's and master's degrees in Electrical Engineering from Virginia Tech and Cornell University, respectively, and has more than 20 years experience as a radio frequency engineer.

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The value of ferrite RF isolators

Proper use at crowded communications sites can reduce intermodulation effects

The class C final stage of a transmitter can be a hot breeding ground for intermodulation products. These intermodulation products travel back up the transmission line to the antenna and can wreak havoc on nearby receivers. This is caused by high-level signals entering the final stage through the antenna and transmission line.

Attenuating these high-level signals before they reach the mixing point (class C final stage) can greatly reduce the level of the intermodulation product(s) formed there. By attenuating the mixing components before they reach the mixing point, a great advantage or leverage is obtained.

Using ferrite radio frequency (RF) isolators between the transmitter and the antenna is good practice. Many times it is possible to manipulate the tuned frequency of the isolator in order

to achieve better results in suppressing intermodulation products that are formed within the final class C stage of the transmitter. Figure 1 shows a typical arrangement using a ferrite RF isolator placed between the transmitter and the transmission line. This particular isolator is a single-section isolator.

Referring to Figure 1, the output of the transmitter is connected to the input port (left side) of the isolator. The output port (right side) is connected, through a second harmonic filter, to the transmission line. The load port (bottom) is connected to the dummy load. Fairly strong second harmonics can be generated within the ferrite isolator. Therefore, the second harmonic filter is used to suppress this second harmonic. It usually is unnecessary to use the special second harmonic filter when bandpass filters are employed at the transmitter output.

In the forward direction, the isolator is very broadband and causes very little attenuation to the transmitter output signal. For example, a typical single-section isolator of the type shown in Figure 1 might offer as little as 0.25 dB of insertion loss in the forward direction. This means that 94% to 95% of the power applied to the input of the isolator appears at the output. For example, 100 W at the input to the isolator produces 94 W to 95 W at the output.

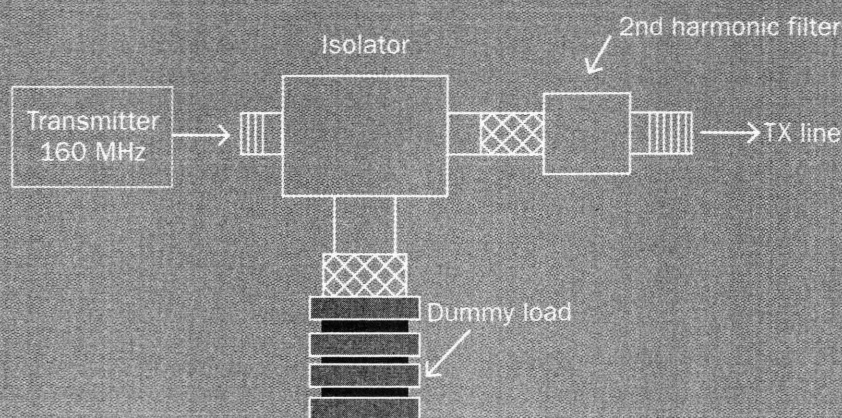
Figure 2 on page 41 represents a printout from a spectrum analyzer that shows the response of the isolator in the reverse direction. That is, the swept signal is applied to the output port, and the response is taken from the input port. As shown, the center frequency is set to 160 MHz, and the sweep span is 40 MHz, or ± 20 MHz. This sets the range from 140 MHz to 180 MHz. The marker is set to approximately 156 MHz (one division to the left of center). The amplitude or vertical scale is set to 10 dB per division. The low point of the notch (at 160 MHz) is approximately 40 dB down, indicating an isolation of 40 dB at 160 MHz.

Now, suppose that there is a transmitter operating nearby at a frequency of 156 MHz (frequency A) and another transmitter nearby operating at 157 MHz (frequency B). Further suppose that there is a receiver nearby operating on a frequency of 155 MHz (frequency C). The mathematical relationship between these frequencies can be expressed by the following equation:

$$2A - B = C$$

Plugging the numbers

Figure 1



This is a typical setup using a ferrite RF isolator. Since ferrite isolators tend to generate rather strong second harmonics, a second harmonic filter is usually placed on the output of the isolator. See text for more details.

from our example into this equation yields the following result:

$$2 \times 156 - 157 = 155 \text{ MHz}$$

The reader will recognize this as a third-order intermod product. Let's assume that without the isolator in the line that this third-order intermod signal is at -20 dBm at the antenna. With the isolator in the line and with the response shown in Figure 2, the 156 MHz signal is attenuated by approximately 23 dB, and the 157 MHz signal would be attenuated by approximately 25 dB. The total attenuation to the intermod signal can be expressed by the following:

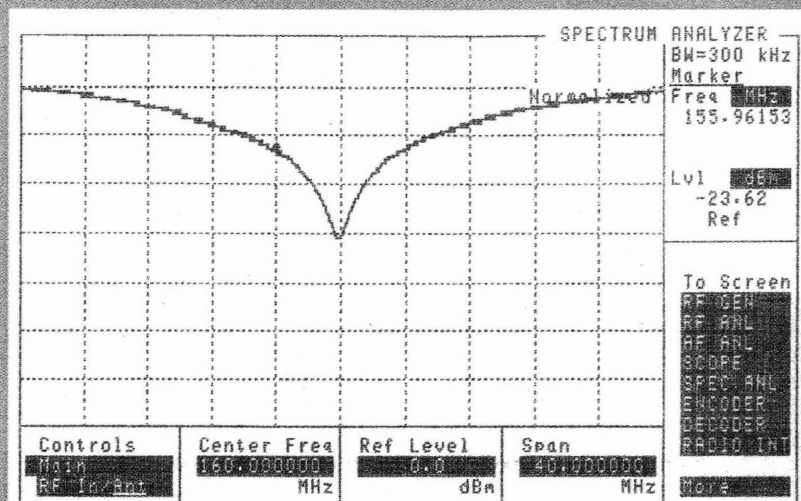
$$2 \times 23 + 25 = 71 \text{ dB}$$

Thus, the intermod signal at C is reduced to:

$$-20 - 71 = -91 \text{ dBm}$$

It is vital to remember the leverage factor. The factor 2A found in the equation on page 40 represents leverage of 2 times the attenuation of signal A. And, since the coefficient of B is 1, we only get

Figure 2



This printout shows the response of the isolator in the reverse direction. The isolation at 160 MHz is greater than 40 dB. The isolation at 156 MHz is about 23 dB.

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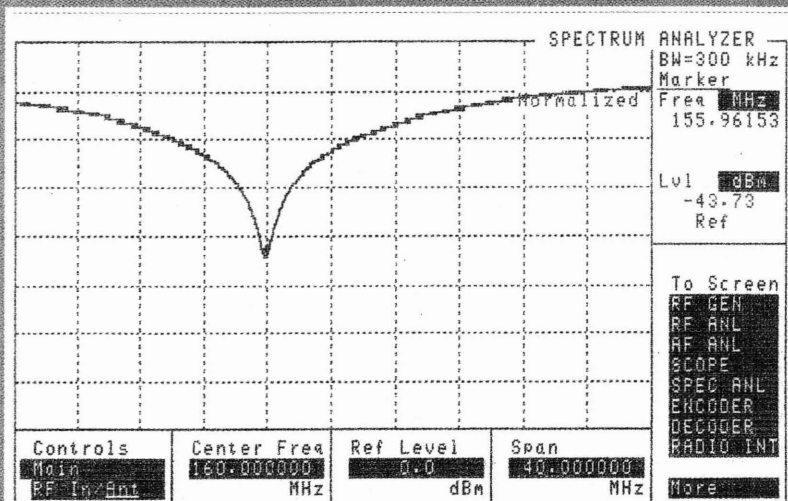
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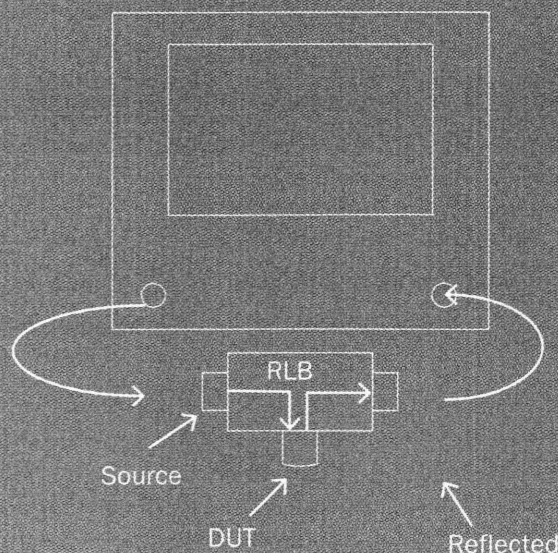
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Figure 3



This printout shows the response of the isolator after being retuned to 156 MHz. Notice that the isolation at 156 MHz is now better than 43 dB. And, the isolation at 157 MHz is approximately 33 dB.

Figure 4



Before the return loss bridge is used, the normalization procedure should be performed. Using this test setup and a shorted connector connected to the DUT port, the response is normalized to 0 dB. This nulls out the losses of the cables and the losses in the bridge itself.

actual attenuation equal to signal B times 1. How can we use this leverage factor to achieve maximum attenuation of the intermod signal on frequency C? We retune the isolator to the frequency with the highest coefficient to get maximum leverage and thus maximum attenuation of the intermod signal.

Figure 3 shows the response of the isolator after it has been retuned for maximum isolation at 156 MHz. As seen from the response curve in Figure 3, the attenuation of signal A (156 MHz) is approximately 43 dB, and the attenuation of signal B (157 MHz) is approximately 33 dB. The total attenuation of the intermod signal at frequency C (155 MHz) is expressed by the following equation:

$$2 \times 43 + 33 = 119 \text{ dB}$$

Thus, the intermod signal at C is reduced to a level of:

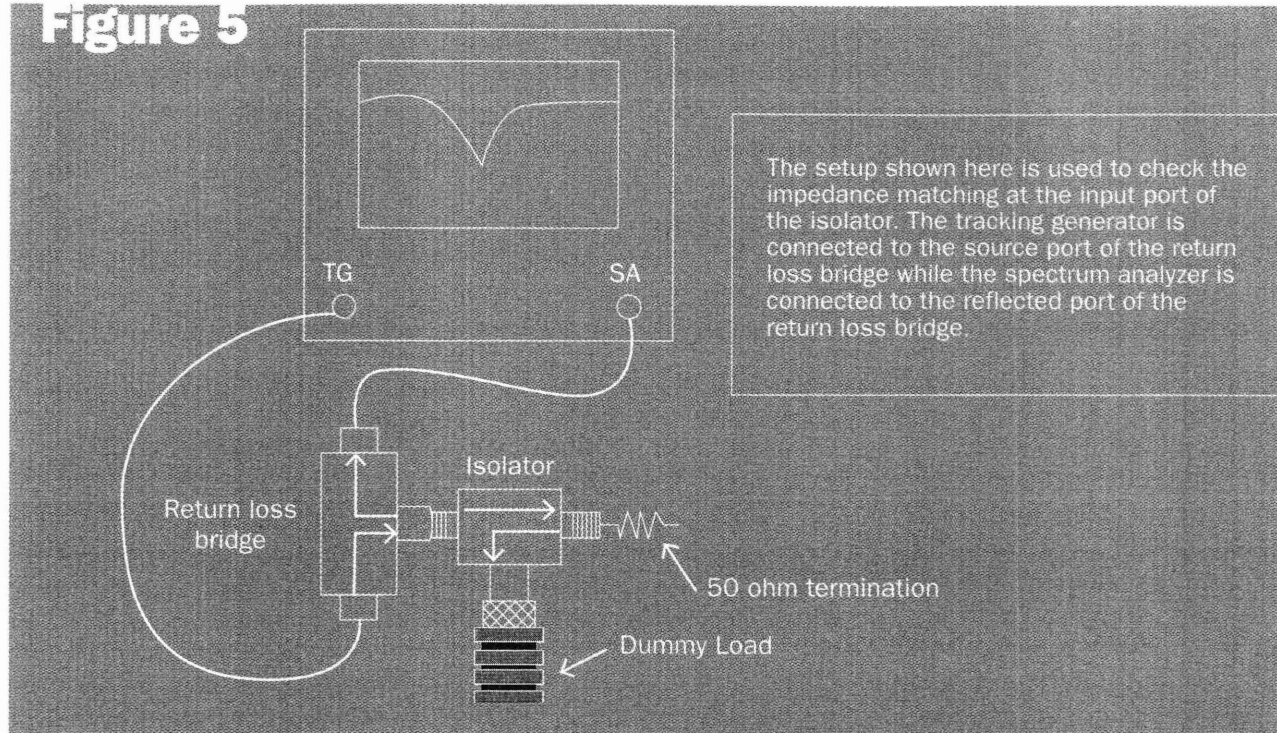
$$-20 - 119 = -139 \text{ dBm}$$

At a level of -139 dBm, the intermod signal is unlikely to cause any harm.

Another important test is to determine the impedance match at the input and output ports of the isolator. A return loss bridge is used for this test. First, the return loss bridge and the interconnections must be normalized. The normalization procedure is performed using the test setup in Figure 4. The test setup for checking the impedance match at the input port of the isolator is shown in Figure 5 on page 43, while the return loss at the input port is shown in the printout in Figure 6 on page 44. The return loss at 160 MHz is more than 50 dB. This indicates a VSWR of approximately 1.006:1, indicating an excellent match.

The same procedure is used to check the impedance match at the output port. The connections to the input and output ports are simply reversed from that shown in Figure 5. The printout for this procedure is depicted in Figure 7 on page 44. The return loss at the output port is greater than 45 dB at 160 MHz. This represents a VSWR of approximately 1.01:1—again, an excellent match.

Figure 5



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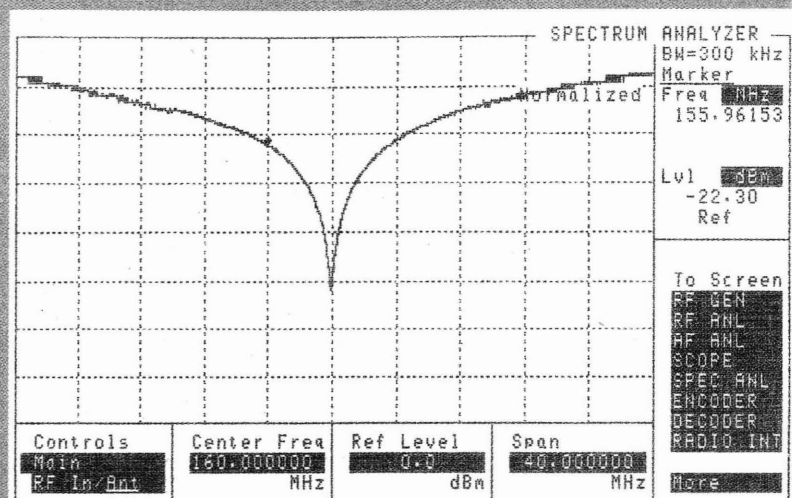
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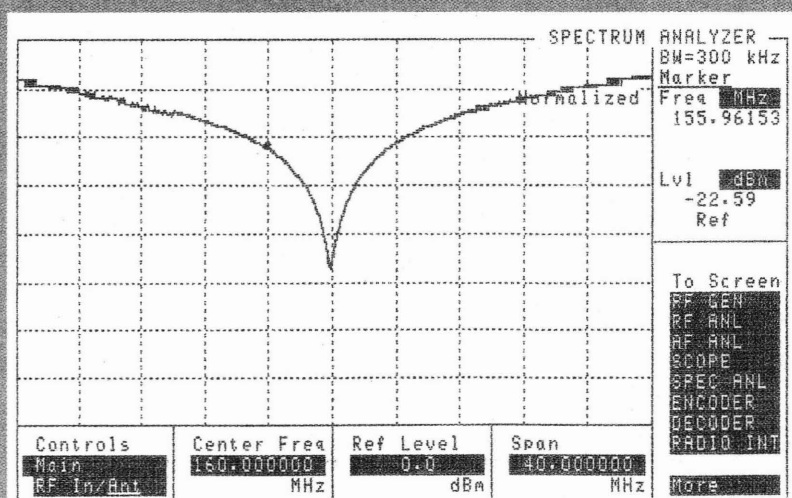
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Figure 6



This printout shows the return loss at the input port of the isolator. The return loss is better than 50 dB, indicating a VSWR of approximately 1.006:1—an excellent match.

Figure 7



The return loss at the output port of the isolator is performed in the same manner as described for the input port—the connections are simply reversed. This printout shows that the return loss at the output port is greater than 45 dB. This represents a VSWR of approximately 1.01:1—also an excellent match.

Now, let's look at a situation where the transmitter output is 100 W at 160 MHz, and the reflected power appearing at the output port of the isolator is 20 W. Referring to the printout in Figure 3 on page 42, the isolation at 160 MHz is about 22 dB. This means that the RF power coming into the isolator's output port (20 W) is attenuated by 22 dB as seen at the transmitter output. Thus, the reflected power appearing at the transmitter output is 20 W attenuated by 22 dB, or 0.13 W.

With a forward power of 100 W and reflected power of 0.13 W, the VSWR at the transmitter output is less than 1.1:1. If the forward power at the output port of the isolator is 95 W, and the reflected power is 20 W, then the VSWR at that point is around 2.7:1. Thus, the isolator is doing an excellent job of maintaining the proper impedance at the transmitter output.

Some important points to observe when using ferrite RF isolators are:

- Observe the manufacturer's recommendation concerning the physical mounting of the isolator.

- Make sure the isolator is properly tuned.

- Use proper 50 ohm load at the load port (usually supplied by the manufacturer).

- Check forward power at the input and output ports of the isolator.

- Calculate the insertion loss from the forward power measurements.

- Check insertion loss against the manufacturer's specifications.

- Never place an isolator in the receive leg.

- The dummy load at the load port should be able to handle at least 50% of the transmitter power.

- Make sure that the isolator is not over-heating (this can lead to destruction).

- Use a second harmonic filter when appropriate.

When properly used, isolators can significantly reduce interference problems at crowded communication sites.

Until next time—*stay tuned!* ■

Diversity is a good thing

Combining multiple receiver antennas improves performance in multipath fading

Antenna diversity is a method for combining multiple, independent, fading signals in a diversity receiver. It is a powerful technique for improving receiver performance in the presence of multipath fading. Interestingly, antenna diversity is used almost universally at cellular sites, but it is rarely used in public-safety or private land mobile radio systems.

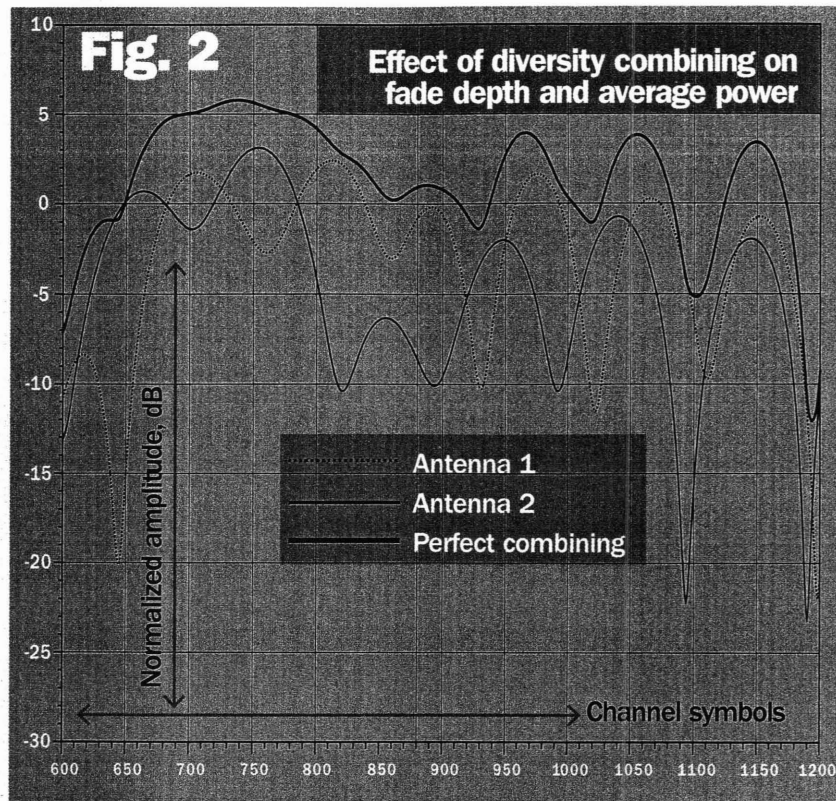
At a high level, the advantage of antenna diversity is simple—two signals are better than one. Figure 1 illustrates this basic concept.

In theory, diversity receivers can combine signals from any number of antennas, but in practice, it is rare to combine more than two antennas. At a minimum, two antennas combined optimally will double the average received power and provide a 3 dB improvement in link performance. But if the signals from each antenna are independent and arrive via a Rayleigh fading channel, the potential improvement is much greater than 3 dB, as we will demonstrate shortly [1].

There are two popular types of antenna diversity for mobile radio channels:

- Space diversity.
- Polarization diversity.

In each case, the objective is to cap-

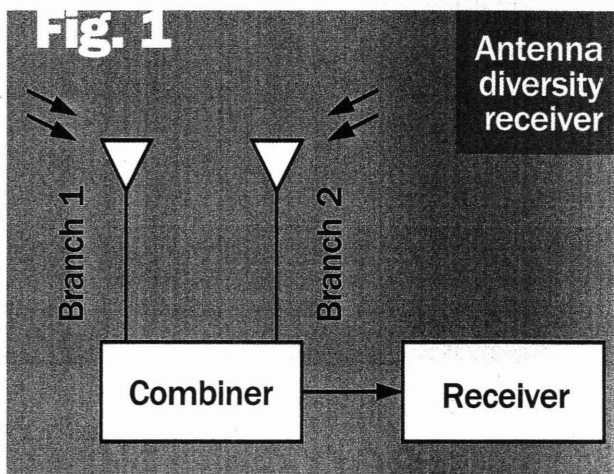


ture multiple independent (or nearly independent) signals and combine them in an optimal manner. With space diversity, antennas are separated by some horizontal or vertical distance that is sufficient to de-correlate arriving signals. Point-to-point microwave links often use vertical separation, but early experiments on LMR channels showed that horizontal separation was more effective because of the nature of scattering on the mobile radio channel [1]. Space diversity has been used on mobile

radio channels since at least the 1970s.

A signal that originally was transmitted with a fixed polarization (usually vertical) does not arrive at the receiver with the same polarization because objects in the path reflect and diffract (bend) the signal and cause the polarization to rotate. In fact, energy at the receiver is spread unequally over many polarization angles from vertical to horizontal.

Polarization diversity, a relative newcomer, exploits this phenomenon of polarization rotation. It is common in the cellular radio industry to employ 45° "slant" polarization, using two antenna arrays in one radome with the antenna elements orthogonal to each other and at a 45° angle to vertical. One of the advantages of polarization diversity is that it limits the number of required antenna assemblies, reducing



the wind load on the tower and making zoning approval more likely.

There also are two popular methods for combining the outputs of diversity antennas:

- Switched combining.
- Maximal ratio combining.

Combining is a misnomer for switching combiners because only one branch is selected at a time. A switching combiner samples each antenna output and selects the path with the best signal, usually the one with the greatest amplitude, but signal-to-noise ratio and other metrics also are used. Switching combiners improve the signal quality in multipath fading, but they are not optimal.

In contrast, maximal ratio combining performs an optimal weighting of the individual signals and combines all diversity signals into one output. It is the optimal form of combining [1].

Despite the advantages of maximal ratio combining, switched combining is still very popular, especially with time division duplexed (TDD) antenna systems that transmit and receive on the same frequency and on the same antenna, but at different time slots. TDD combiner systems attempt to pick the optimal antenna for receiving and transmitting.

Diversity antenna systems are most effective in the presence of multipath fading. By selecting the best antenna signal (switched combiner) or combining the antenna outputs (maximal ratio combining), the net effect is to reduce the depth of fades. This effect is shown in Figure 2, where we have plotted two independent Rayleigh fading channels and their combined output, assuming maximal ratio combining.

We now know that diversity combining reduces the fade depth on a Rayleigh fading channel and can increase the average power by a factor proportional to the number of diversity antennas. But we also want to know how this improvement is put to use in the receiver and exactly how much improvement should

be expected in a modern digital radio.

Using the Rayleigh fading channel model, we can derive expressions for error probability for the non-diversity and two-branch diversity cases. We'll assume binary frequency shift keying (FSK), and note that the derivations for C4FM and CQPSK, the Project 25 modulation techniques, are similar.

On the static Gaussian noise channel, the probability of bit error for binary FSK is given by Equation 1 [2, pp. 717].

One can show that on a Rayleigh fading channel, E_b/N_0 is exponentially distributed. If we calculate the mean bit-error rate over the exponential probability density function, we get the probability of bit error shown in Equation 2 [2, pp. 717].

Note from Equation 2 that instead of decreasing exponentially with signal-to-noise ratio, now the error rate decreases only inversely with signal-to-noise ratio. Consequently, the required E_b/N_0 at a bit-error rate of 10^{-5} is 37 dB higher in Rayleigh fading (Figure 3 on page 69). So we incur a tremendous

mitter power, but increasing the power of a battery-operated radio by 37 dB is far from practical—a 3 W radio would turn into a 15,000 W radio. Diversity is a more practical technique. The probability of bit error for an L-branch diversity combiner with FSK modulation is found in many textbooks [2, pp. 723-726].

An important principle is that the probability of bit error for an L-branch diversity system decreases as the inverse of the signal-to-noise ratio raised to the Lth power. For example, for L=2 branch diversity, the probability of bit error decreases according to the inverse squared signal-to-noise ratio. This behavior is shown in Figure 3, where we have plotted the probability of bit error for a two-branch FSK diversity receiver.

Note from Figure 3 the large improvement in required E_b/N_0 for a bit error rate of 10^{-5} . However, we are still 17.3 dB shy of non-fading performance. More diversity antennas would help, but most systems stop at two antennas for cost reasons and try to make up the

Equation 1

$$\text{Probability of bit error (FSK)} = 1/2 \exp(-snr/2)$$

where snr is the ratio of energy per bit to noise power spectral density, also denoted by E_b/N_0 .

Equation 2

$$\text{Mean probability of bit error in Rayleigh fading (FSK)} = 1/(2 + \langle snr \rangle)$$

where $\langle snr \rangle$ is the mean value of E_b/N_0 in Rayleigh fading.

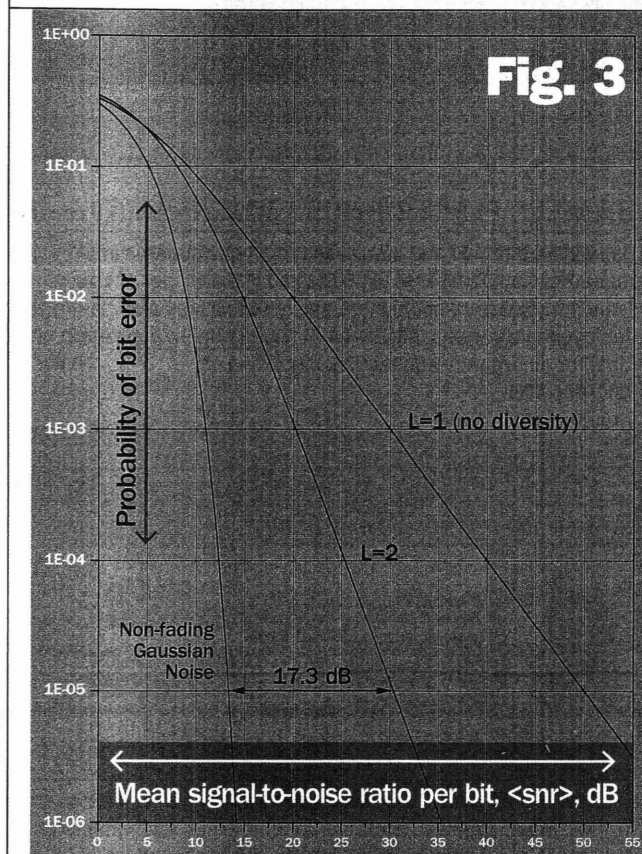
power penalty simply by operating in a fading environment.

The obvious way to recover this lost performance is to increase the trans-

mission power, but increasing the power of a battery-operated radio by 37 dB is far from practical—a 3 W radio would turn into a 15,000 W radio. Diversity is a more practical technique.

Diversity systems normally operate

Continued on page 69



Continued from page 45

from the cellular base station or, in rare cases, from the two-way radio repeater site. Diversity also works at the mobile receiver. Smaller antenna separations are required for space diversity at the mobile receiver because the scatterers are relatively close to the receiver. Mobile diversity antenna systems are not common today for cost and space reasons, but several companies are working on mobile diversity systems, and IEEE 802.11 radios may be the first to use them widely. ■

Jay Jacobsmeyer is president of Pericle Communications Co., a consulting engineering firm located in Colorado Springs, Colo. He holds bachelor's and master's degrees in Electrical Engineering from Virginia Tech and Cornell University, respectively, and has more than 20 years experience as a radio frequency engineer.

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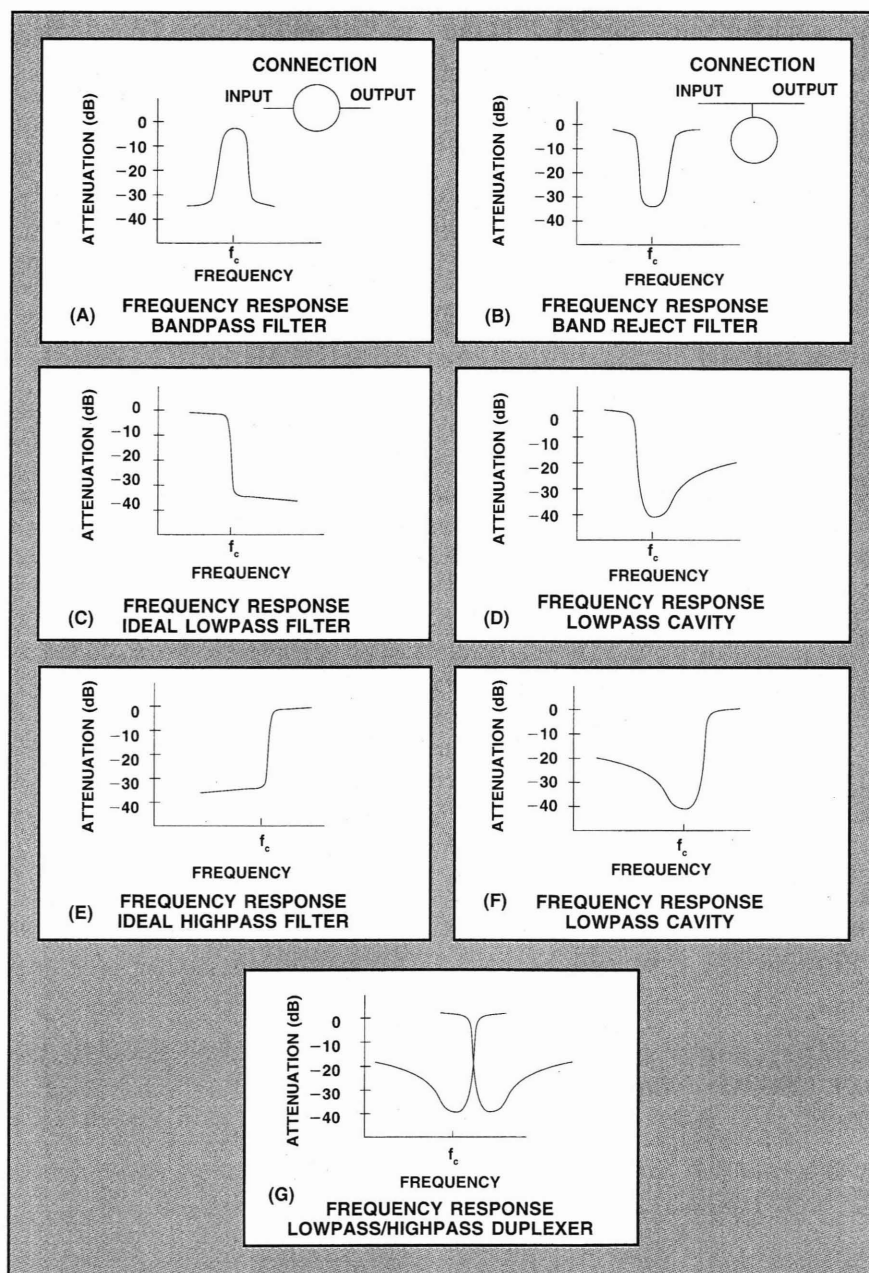
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How to use duplexers: Installing and tuning

Part 3—How duplexers are constructed and how they are installed have much to do with their tuning and frequency stability. Here are tips for ordering duplexers and for initial tuning or retuning.



By Brian J. Henderson, P. Eng.

Whether duplexers are used in a typical manner to support repeater operation or as filters to reduce or eliminate interference, their construction, installation and tuning affect the outcome.

Temperature

Temperature plays a big part in duplexer performance.

The volume of the cavity is extremely critical to duplexer tuning. A small temperature change can contract or expand cavity components, shifting the duplexer frequency as much as 30kHz–40kHz, enough to shut down a repeater system.

The system cannot be allowed to shut down because of a duplexer temperature problem. What do you do?

The first thing is to make sure the temperature in your repeater building remains relatively constant. Install a thermostat on an electric heater to heat a small building. Keep duplexers themselves out of drafts and away from ventilation fans and windows. Keep them out of sunlight so they are not subject to heating during the day and cooling at night.

Physical construction of duplexers

There are two methods of duplexer construction.

Figure 1. The easiest way to understand the use of a spectrum analyzer and tracking generator for cavity tuning is by example. Attenuation vs. frequency curves for bandpass and band-reject cavities are shown in (A) through (G) to the left.

Henderson is senior engineer, communications, with Canadian Western Natural Gas, Calgary, Alberta.

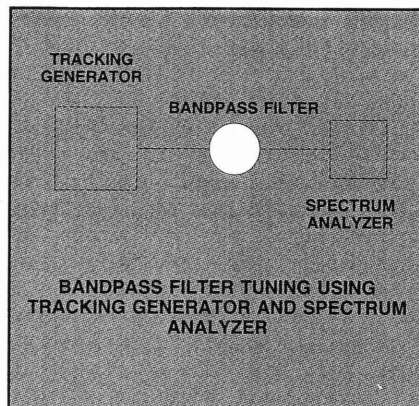


Figure 2. To replicate the example of cavity tuning described in the text, connect a tracking generator and spectrum analyzer as shown.

The older method uses round containers and associated internal components. These duplexers are more difficult to mount on equipment racks because of their shape.

Manufacturers then developed rectangular duplexers. They are compact and far more convenient for rack-mounting.

✱ When round cavities are exposed to temperature changes, they expand equally in all directions. Tuning may or may not be affected.

✱ When a rectangular cavity changes temperature, it expands more in one direction than another, which is certain to affect its tuning.

Generally speaking, where temperature extremes are expected, such as for unheated mountaintop locations, use a round cavity filter.

Coaxial cable

There are several factors to consider when using coaxial cable to connect cavities to make a duplexer.

(1) To preserve as much isolation as possible, always use double-shielded cable.

Use either RG-214/U or RG-400/U cable for all connections. It is a good idea to use these types of cable as jumper cables to the radio repeater and antenna connections. This keeps RF signals where they should be—inside the radio and antennas.

(2) Cable length is important.

✱ Do not use random-length jumper cables. All cable lengths are to be either a quarter-wavelength ($1/4\lambda$) or a half-wavelength ($1/2\lambda$) to maintain the impedance presented by a cavity to the line.

(3) When calculating the cable length, take into account the cable's velocity factor.

When RF travels in a cable, it slows down. The difference between its velocity in free space or air and its velocity

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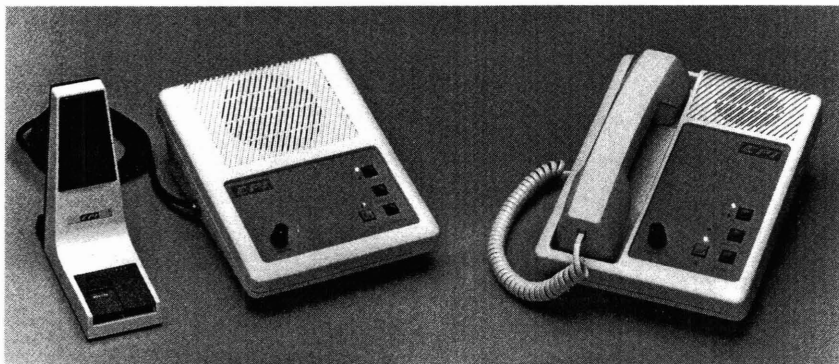
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in cable is the cable's velocity factor, expressed as a percentage.

Cable impedance and length

The coaxial cable jumper length affects the impedance at each end of the cable. Half-wavelength cables have an identical impedance at each end; quarter-wavelength cables have an opposite impedance.

For example, if the impedance at one end of a $\frac{1}{2}\lambda$ cable is high (band-reject), the impedance at the other end will be high.

If the impedance at one end of a $\frac{1}{4}\lambda$ cable is high, the impedance at the other end will be low.

Remember these rules when moving or separating cavities on an equipment rack and changing their connecting cable lengths.

For band-reject cavities, use $\frac{1}{4}\lambda$ cables. Bandpass cavities and cables between radio and filters should be $\frac{1}{2}\lambda$ long.

The spectrum analyzer

A spectrum analyzer with a tracking generator can be used for tuning duplexers.

A spectrum analyzer is similar to an oscilloscope. It displays frequency on the horizontal axis vs. amplitude on the vertical axis. Because duplexers are frequency-and-amplitude-dependent, the spectrum analyzer is ideal for tuning and aligning them.

The easiest way to understand the use of a spectrum analyzer and tracking generator is by example. Attenuation vs. frequency curves for bandpass and band-reject cavities are shown in Figure 1 on page 24.

As the name implies, a bandpass cavity has low attenuation at its bandpass frequency. Other frequencies are attenuated.

Duplexer series

Previous installments in this article series include:

- "How To Use Duplexers: Isolation Requirements" in the July 1994 issue.
- "How To Use Duplexers: The Various Types" in the August 1994 issue.

Back issues printed within the past two years can be ordered for \$5 each, postpaid. Call customer service at 800-441-0294. Copies of articles printed more than two years ago are unavailable from the publisher.

The farther from the cavity's center frequency, the higher the attenuation.

A band-reject cavity is the opposite. A single frequency is rejected or "shorted out" of the circuit. Maximum attenuation occurs at the reject frequency. The further from the cavity's center frequency, the lower the attenuation.

Filter tuning

How can this information be used for

tuning?

Remember the cavity configuration—*bandpass* or *band-reject*. Use the tracking generator and analyzer to produce the correct frequency curve as shown on the spectrum analyzer.

For example, consider a bandpass cavity to be tuned to a center frequency of 153.620MHz. Set the spectrum analyzer for a center display frequency of 153.620MHz. Connect the tracking gen-

erator and spectrum analyzer as shown in Figure 2 on page 27.

Loosen the locking set screws or nuts on the cavity's tuning rod. Slide the rod in or out so that the attenuation is minimized around the analyzer display's center frequency. The cavity is now tuned. Figure 1A shows the display that should appear on the analyzer.

A reject cavity produces a reverse display. Again set the spectrum analyzer center frequency to, for example, 153.620MHz. Adjust the tuning rod so that the rejection notch on the display is centered and on the correct frequency. (See Figure 1B.)

If the cavity is a band pass/band-reject type, there are two adjustments to be made. Example frequencies are 153.62MHz (pass) and 154.22MHz (reject). The bandpass adjustment is set first using the cavity tuning rod. The band-reject is set by adjusting the capacitor.

The capacitor usually is right beside the input connector on the cavity itself. It must be adjusted using either a nylon or other non-inductive tuning tool. (See Figure 1D and 1F for lowpass and highpass response curves.)

Note that most highpass/lowpass cavities can be adjusted for either highpass or lowpass configurations within certain limits by watching the analyzer display while adjusting the capacitor.

The spectrum analyzer must be used to start the tuning process, and it gets you near the correct adjustments. The analyzer will indicate whether a cavity is set for highpass or lowpass.

At the factory, fine tuning is performed with a network analyzer and return loss bridge. When such equipment is unavailable, another fine-tuning method can be used with a transmitter, wattmeter and signal generator. The filters can then be put into service with a reasonable confidence level.

Tuning rods too short?

When tuning a filter, you may find that you "run out" of tuning rod.

To go lower in frequency, the tuning rod must be pushed into the cavity. You may reach a point where the end of the rod outside the cavity is too short to push it any farther, so the cavity cannot be adjusted any lower in frequency.

When a manufacturer builds a duplexer, all the rods initially are the same length. Depending on the frequency requested when the unit originally was ordered, some tuning rods may appear longer than others after tuning.

Most manufacturers then cut the rods so that they all protrude the same distance outside the cavity. This cutting is

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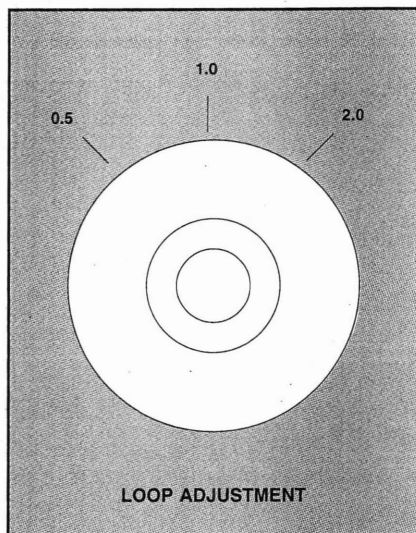


Figure 3. Inside the cavity, a loop connected to the cable connector forms the radiating and pickup element. The loop's position can change the cavity's performance. In newer cavities, the loop may be marked 0.5, 1.0 or 2.0. The loop turns when certain screws are removed. Turning the loop makes the cavity's frequency response sharper or wider as required. If an extremely selective cavity is required, rotate the loop while watching the spectrum analyzer display to narrow the cavity's bandwidth. See text for precautions.

for aesthetic reasons only. Cavities look better when their rods are an equal length—that is, if the cavity never is retuned.

Unfortunately, a cavity supplied with a cut rod may cause problems during tuning, especially when its operating frequency is moved much—from 165MHz to 151MHz, for example.

You have two options: Try to find a cavity with a longer tuning rod, or buy a new rod from the duplexer manufacturer, which can be expensive.

When ordering filters, it is a good idea to specify on your purchase order: "Tuning rods not to be cut." That way, if the frequency ever has to be changed, options are left open for further tuning or adjustment.

Tuning loops

Inside the cavity is a loop connected to the cable connector that forms the radiating and pickup element.

The loop's position can change the cavity's performance. In newer cavities, the loop may be marked 0.5, 1.0 or 2.0. The loop turns when certain screws are removed. (See Figure 3 to the left.)

Turning the loop makes the cavity's fre-

quency response sharper or wider as required. If an extremely selective cavity is required, rotate the loop while watching the spectrum analyzer display to narrow the cavity's bandwidth.

There is a trade-off. If the loop is rotated for a narrower passband, for example, insertion loss increases. Numbers on the cavity refer to its insertion loss measured in decibels. Note that if the cavity is a bandpass type, both loops must be rotated to, for example, the 1.0dB position. Otherwise, loss will increase substantially. The cavity must remain symmetrical for proper operation.

Loop size

Tuning loops themselves are a particular length, depending on frequency.

Making a big change in frequency may require changing the loop size as determined by the manufacturer. When new loops are required, contact the manufacturer for parts and components.

Next: After initially tuning a filter using a spectrum analyzer and the technique described above, steps described in Part 4 can be used with inexpensive test equipment for fine-tuning.



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How do you spell VSWR?

By Harold Kinley, CET

Of course, there is only one way to spell VSWR, but there are many ways to represent antenna (or other RF load or source) mismatches using different

measuring units.

For example, occasionally you will see the terms *return loss*, *reflection coefficient*, *percent reflected power* and *reflected-to-forward power ratio*. It is enough to confuse Confucius!

But do not despair. There is a way to

convert among the various units without using calculus.

► **VSWR**—Probably the most widely used and best understood measuring unit used by radio communications technicians is voltage standing wave ratio (VSWR).

When a manufacturer specifies that a certain antenna operates over a certain frequency range at a VSWR of 1.5:1 or better (less), we immediately know that this is a good match—the system will operate efficiently with minimal wasted power.

► **Return loss**—When the forward power on a transmission line equals 100W and the reflected power equals 25W, the return loss is determined by the following formula:

$$\begin{aligned} R_L &= 10 \log (P_F / P_R) \\ &= 10 \log (100 / 25) \\ &= 10 \log (4) \\ &= 10 \times 0.602 \\ &= 6.02 \text{ dB} \end{aligned}$$

where

R_L = return loss in decibels

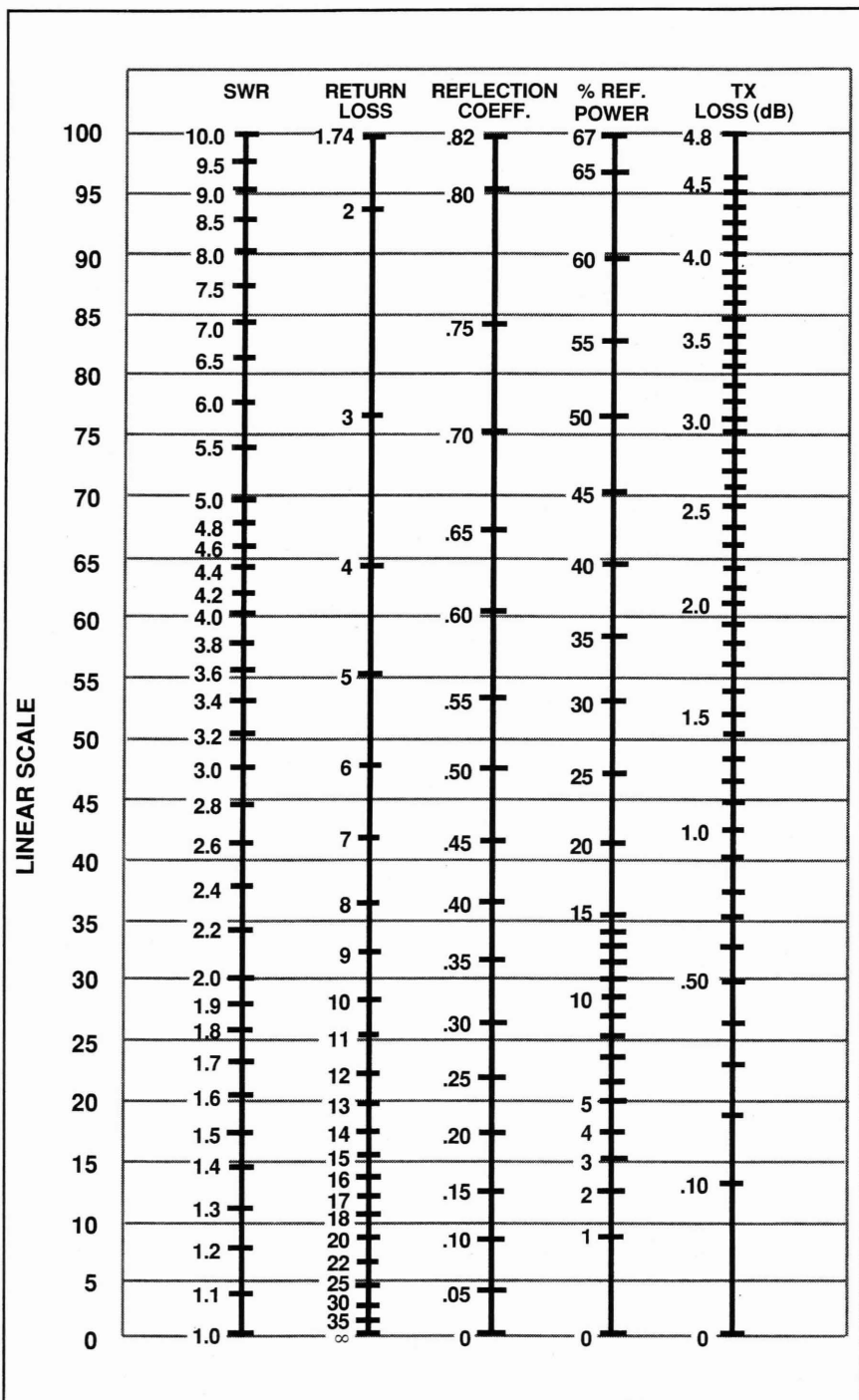
P_F = forward power in watts

P_R = reflected power in watts

Because of transmission line loss, the VSWR at the antenna is worse than what is measured at the transmitter, and the same difference applies for every unit of measure.

It is easy to convert return loss at the transmitter to return loss at the antenna because the transmission line loss usually is stated in terms of decibels per 100 feet (dB/100ft.).

For example, if the line loss is 2dB and the return loss at the transmitter is (continued on page 58)



Kinley is a certified electronics technician with the South Carolina Forestry Commission, Spartanburg, SC. He is the author of *Standard Radio Communications Manual: With Instrumentation and Testing Techniques*, Prentice-Hall, 1985.

Technically speaking

(continued from page 8)

10dB, then the return loss at the antenna is:

$$\begin{aligned} R_L &= 2(L_L) \\ &= 10 - 2(2) \\ &= 6\text{dB} \end{aligned}$$

where

R_L = return loss in decibels at the transmitter

L_L = transmission line loss

It is necessary to multiply the line loss by a factor of 2 because both the forward power and the reflected power are attenuated by the line loss. In this example, the forward power is 2dB lower at the antenna, and the reflected power is 2dB higher at the antenna.

► **Reflection coefficient**—Simply stated, reflection coefficient is the ratio of

voltage of the incident wave to voltage of the reflected wave.

Because voltage is proportional to the square root of the power, reflection can be defined as $\sqrt{(P_R)/\sqrt{(P_F)}}$ or simply $\sqrt{(P_R/P_F)}$.

For example, if the forward power is 100W and the reflected power is 25W, then the reflection coefficient Γ is calculated as follows:

$$\begin{aligned} \Gamma &= \sqrt{(25/100)} \\ &= \sqrt{(0.25)} \\ &= 0.5 \end{aligned}$$

► **Reflected-to-forward power ratio**—The reflected-to-forward power ratio unit of measure probably is the most familiar to land mobile radio technicians because they favor the use of an in-line directional wattmeter to measure forward and reflected power on the trans-

mission line.

It is simple, then, just to compare the reflected power reading with the forward power reading to get a quick mental picture of the degree of mismatch on the line.

For example, if the forward power is 100W and the reflected power is 25W, then the ratio (reflected-to-forward) is 0.25. Notice that this figure is the *square of the reflection coefficient* that was obtained using the same reflected and forward power measurements.

► **Percent reflected power**—The percent reflected power figure is obtained by multiplying the reflected-to-forward power ratio by 100.

For example, a ratio of 0.25 equals 25% reflected power.

Transmission loss—Transmission loss is the actual loss in the transmitted signal resulting from the mismatch.

For example, with 100W of forward power at the antenna and 25W of reflected power, only 75% of the *available power* is transmitted. The transmission loss is:



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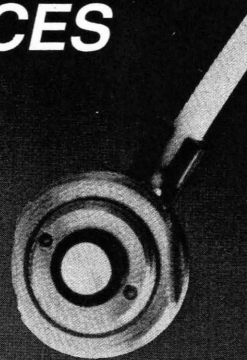
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Technically speaking

$$T_L$$

$$= 10 \log(1 - r)$$

$$= 10 \log(1 - 0.25)$$

$$= 10 \log(0.75)$$

$$= 10 \times -0.125$$

$$= -1.25 \text{ dB}$$

Conversion formulae: The following formulae can be used to convert between any of the measuring units discussed.

$$S = \frac{1 + \sqrt{r}}{1 - \sqrt{r}}$$

$$S = \frac{1 + \Gamma}{1 - \Gamma}$$

$$S = \frac{1 + 10^{-(R_L/20)}}{1 - 10^{-(R_L/20)}}$$

$$r = \left(\frac{S-1}{S+1} \right)^2$$

$$r = \Gamma^2$$

$$r = 10^{-(R_L/10)}$$

$$\Gamma = \sqrt{r}$$

$$\Gamma = \frac{S-1}{S+1}$$

$$\Gamma = 10^{-(R_L/20)}$$

$$R_L = 10 \log \left[\left(\frac{S-1}{S+1} \right)^2 \right]$$

$$R_L = 10 \log \left(\frac{1}{r} \right)$$

$$R_L = 10 \log \left(\frac{1}{\Gamma^2} \right)$$

$$T_L = 10 \log \left[1 - \left(\frac{S-1}{S+1} \right)^2 \right]$$

$$T_L = 10 \log(1 - \Gamma^2)$$

$$T_L = 10 \log(1 - r)$$

$$R = 100r$$

where

$$S = \text{VSWR}$$

$$\Gamma = \text{reflection coefficient}$$

$$T_L = \text{transmission loss in decibels}$$

$$R_L = \text{return loss in decibels}$$

$$r = \text{reflected-to-forward power ratio}$$

$$R = \text{percent reflected power}$$

Nomograph—The nomograph in Figure 1 on page 8 converts between equivalent units of VSWR, return loss, percent reflected power, reflection coefficient and transmission loss that lie on the same line. At a glance, you can convert from any unit of measure to another.

There are a few advantages that are not so apparent at first glance.

For example, if you know the VSWR at the transmitter and the line loss, you can find the VSWR at the antenna by

using the following example as a guide.

The VSWR at the transmitter is 2.0:1, and the line loss is 2.5dB. Locate 2.0 on the VSWR scale and read the equivalent return loss in decibels on the next scale. It is approximately 9.5dB.

Subtract twice the line loss ($2 \times 2.5 = 5$) from the return loss reading, and move to that point on the return loss scale. This point will be 4.5dB on the return loss scale, which is equivalent to a VSWR of approximately 4:1 at the antenna.

The nomograph can be used to determine the maximum VSWR that can be seen at the transmitter for a given amount of line loss.

For example, suppose the line loss is 3.0dB. Double this figure and move to 6 on the return loss scale. Follow the line out to the VSWR scale, and note that for a 3dB line loss, the maximum VSWR seen at the transmitter is approximately 3.0:1.

The *transmission loss* scale may provide some surprises.

For example, notice that for a VSWR of 3.0:1, the transmission loss is only 1.25dB. It appears, then, that a VSWR of 3.0:1 is not so terrible after all.

Here is the catch: Modern radio transmitters have an automatic power reduction circuit that limits output when a high VSWR is detected. Thus, the practical transmission loss would be much greater for such radios than would be indicated by the nomograph.

For example, if power output is reduced 50% for a VSWR of 3.0:1, this reduction would add 3dB to the *normal* transmission loss to make the practical effect a net 4.25dB transmission loss.



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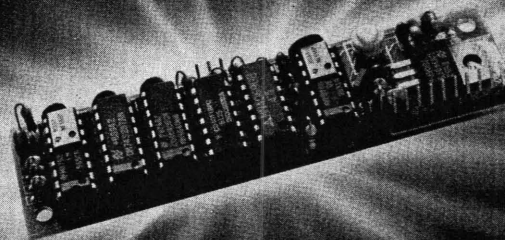
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How to use duplexers: Combiners, multicouplers

Part 5—Transmitter combiners and receiver multicouplers allow multiple repeaters to share the same antenna. If the system can tolerate the energy loss, combiners allow more efficient use of antenna space.

By Brian J. Henderson, P. Eng.

Transmitter combiners and receiver multicouplers are filter configurations.

A duplexer allows a transmitter and receiver to use a single antenna simultaneously. A transmitter combiner allows several transmitters to use the same antenna without interfering with each other; a receiver multicoupler does the same for receivers. Where tower space is at a premium, a combiner and a multicoupler may allow additional radios to use previously installed antennas.

When interference occurs between some transmitters or receivers at a radio site, a combiner and multicoupler system reduces the interference. Interference is common on radio towers with separate antennas for each repeater system. The isolation of a combiner and multicoupler system may be required to prevent intermodulation products (intermod) from occurring.

Combiner systems have losses. A combiner system allows additional transmitters and receivers to use the same antenna.

A combiner system may be warranted if:

- the tower is loaded with antennas.
- interference between systems is occurring.
- the tower must be rebuilt to accommodate more radio systems.

A combiner system may be expensive and it may introduce system losses that affect radio coverage. These factors must be considered before installing a combiner option. On the other hand, using combiners may cost less than rebuilding a tower to accommodate more antennas or radio systems.

Henderson is senior engineer, communications, with Canadian Western Natural Gas, Calgary, Alberta.

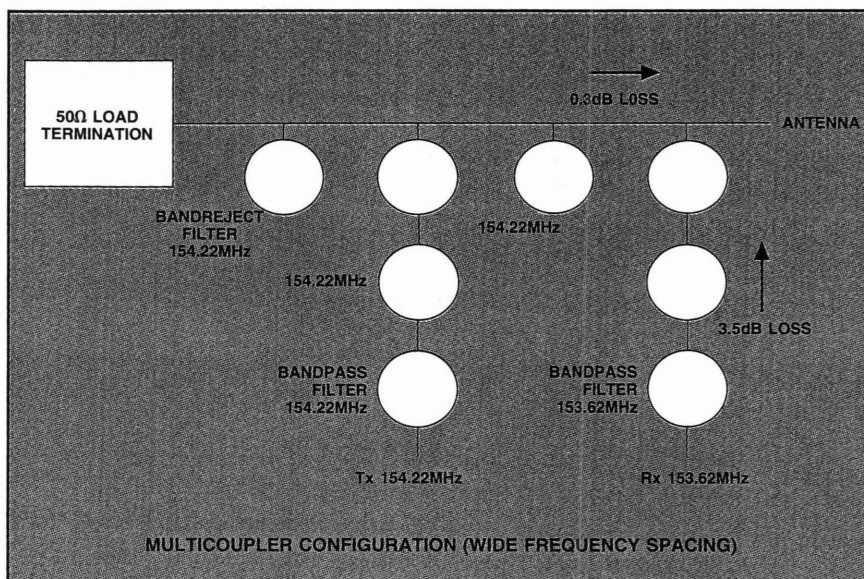


Figure 1. A typical bandpass/band-reject combiner is configured as shown. It includes two or three bandpass cavities and one reject cavity per frequency. Because of the combiner's bandpass nature, the minimum frequency separation is 2MHz. For closer frequency spacing, use the bandpass/isolator type of combiner configuration.

There are two basic combiner types.

Whether to use either depends on transmitter and receiver frequency spacing.

As with duplexers, a combiner's purpose is to provide sufficient isolation between transmitters and receivers so that they operate without mutual interference.

Bandpass/band-reject combiner

Figure 1 above is a general drawing of a bandpass/band-reject combiner.

This combiner type has two or three bandpass cavities and one reject cavity per frequency. Because of the combiner's bandpass nature, the minimum frequency separation is 2MHz. For closer frequency spacing, use the bandpass/isolator type of combiner configuration.

Note as well that each "leg" of the sys-

tem has approximately 3.5dB of loss plus 0.3dB system loss between each connection point and the antenna. That loss is why receivers are placed on the "string" first, followed by the transmitters. The idea is to reduce loss between the antenna and receiver.

Tuning these combiner strings consists of tuning individual bandpass and band-reject cavities. Once tuned to the correct frequency using a transmitter and wattmeter as described above, the combiner can be installed. Remember to terminate all unused ports with a 50Ω load.

The combiner can be expanded to add more transmitters and receivers. Figure 1 shows one section of a combiner string. This section allows one transmitter and one receiver to operate.

Adding another eight individual cavities

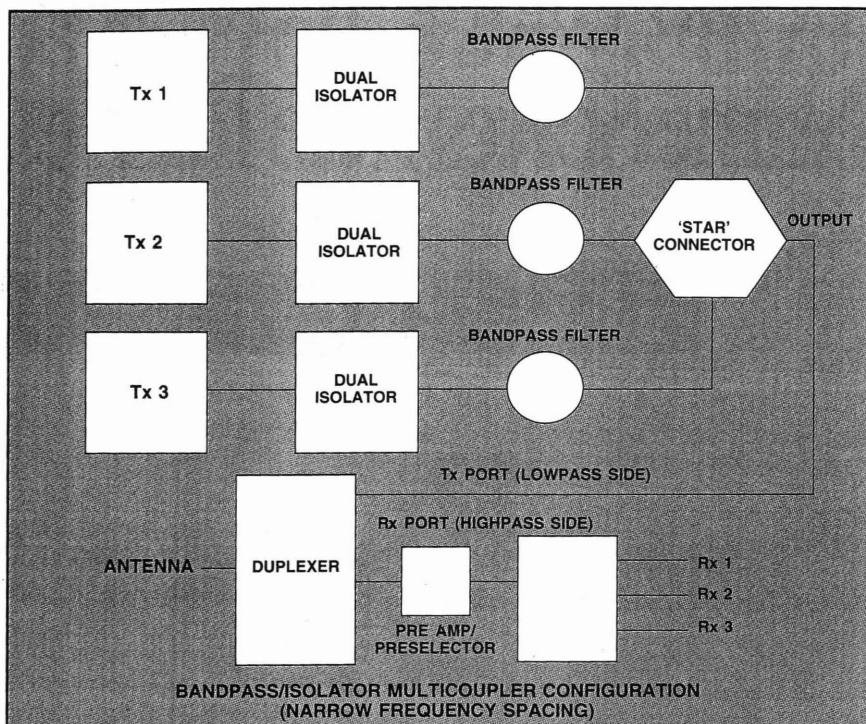


Figure 2. A typical bandpass-isolator combiner is configured as shown. It can be used for frequency spacings closer than 2MHz. An operational limit is a minimum spacing of 500kHz at VHF and 300kHz at UHF. For cellular use, channel spacing as close as 250kHz is possible.

and moving the 50Ω load termination to the rear of the combiner string allows one more transmitter and receiver to be added. Remember to connect the receivers ahead of all transmitters to minimize loss between the antenna and receivers.

The string can be expanded to accommodate as many as four transmitters and four receivers on a single antenna. This is a practical limit because of loss.

The last transmitter in the combiner string (just before the load termination) will experience $8 \times 0.3\text{dB} + 3.5\text{dB} = 5.9\text{dB}$ of loss. More transmitters can be added if the system can tolerate the loss. Normally, four is the maximum number of repeaters used because of the loss.

Bandpass-isolator combiner

The bandpass-isolator type of combiner combination can be used for frequency spacings closer than 2MHz.

An operational limit is a minimum spacing of 500kHz at VHF and 300kHz at UHF. For cellular use, channel spacing as close as 250kHz is possible. Figure 2 to the left is a drawing of a typical bandpass-isolator combiner configuration.

Several frequency restrictions apply to using this combiner type:



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(1) The minimum frequency spacing is 500kHz at VHF and 250kHz at UHF.

(2) All transmit frequencies must fall into the passband on one side of a duplexer.

(3) All receive frequencies must fall into the passband on the other side of the duplexer.

(4) Frequencies cannot be scattered throughout a given VHF or UHF frequency band.

There are four basic components to this type of combiner:

- (1) ferrite isolator.
- (2) bandpass filter.
- (3) duplexer.
- (4) receiver multicoupler.

The duplexer's purpose is to separate the transmitter frequency group from the receive frequency group. The duplexer can be either type, depending on the overall transmitter and receiver frequencies' band-

width requirements.

Transmitters are connected first to a dual isolator, then to a bandpass filter. From there, they are connected to a "star connector," which is similar to a "T" connector except with four or more connection ports. Several cables can be connected in parallel because they represent several circuits connected in parallel with a 50Ω circuit. There is no effect on the impedance at a particular frequency. Note also that unused ports should be left open rather than terminated to preserve isolation and impedance.

Tuning

The bandpass-isolator combiner has only two components that can be tuned in the field.

First, the duplexer can be tuned as described for the type and configuration of the duplexer in use. The bandpass filter can be tuned in the conventional manner as described above for bandpass cavities.

Ferrite isolators should not be adjusted. They usually have enough bandwidth to use without tuning.*

The receiver multicoupler is made of a bandpass network, an amplifier and splitter. The only component that can be adjusted is the bandpass network, if required. The bandwidth usually is quite wide, so adjustment may or may not be necessary depending on how much frequency shift there is among all receivers.

Try injecting signal into a receiver through the receiver multicoupler. If the receiver operates with reasonable sensitivity, leave the bandpass filter alone. If it works, don't fix it.

Duplexer

When combining several transmitters and receivers in this way, it still is necessary to provide separation between the transmitter and receiver groups of frequencies. The easiest way to do this is with another duplexer. Still, make sure that transmitter and receiver frequencies fall into a group that can be separated by a duplexer.

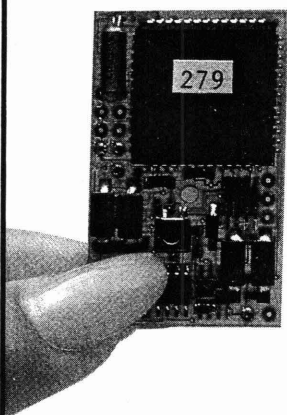
As an example, consider a system with the following operating frequencies:

SYSTEM	TRANSMIT FREQUENCY (MHz)	RECEIVE FREQUENCY (MHz)
1	413.1000	418.1000
2	413.3000	418.3000
3	413.5000	418.5000
4	413.7000	418.7000

In this case, all transmitter frequencies fall into the 413.1000MHz-413.7000MHz

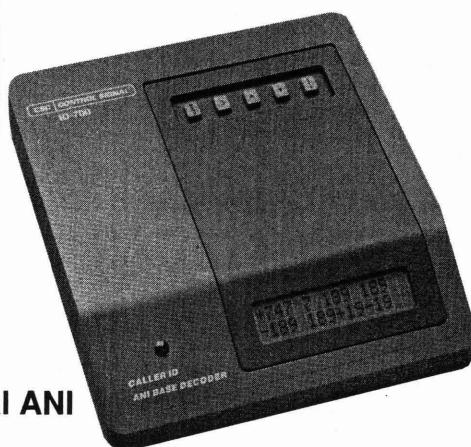
*See "Technically Speaking" by Harold Kinley, C.E.T., in the June and July 1993 issues.

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range. Receive frequencies fall into the 418.1000MHz-418.7000MHz range. A bandpass-style duplexer with a 600kHz passband bandwidth will work fine.

A spectrum analyzer and tracking generator must be used to align this duplexer. Adjust for required passbands on both the transmit and receive sides.

If frequencies are scattered throughout the band, it is not possible to use this multi-coupling system type. There are no other

alternatives besides separate antennas.

Shipping and handling

You have a duplexer or combiner tuned up and ready to travel to the site. You have made sure that all adjustments are locked and tightened. You throw it in the back of the truck, and off you go. You get to the site, install it and find that it does not work. Now what?

Two things can happen to duplexers or

combiners in transit to detune them.

All tuning rods on duplexers have either a set screw or a locking nut to tighten their adjustments. Capacitors do not. They can move if bounced around or jarred. When they move, the cavity is detuned.

The other thing that can happen is that the tuning rod can shift inside the cavity. When it does, the filter is detuned.

The rule is, take it easy with the equipment. It is fragile. If you have the original packing material and shipping crates, use them. Wrap all filters in foam rubber pads to cushion them when driving up the bumpy road to the radio site.

A temperature change may be a factor, too. Filters transported in winter may be cold when they arrive at the site, so let them warm up for 15-20 minutes before installing them. A small change in temperature can detune cavities. Once their temperature stabilizes, the filters should be back into proper tune.

A further note on temperature: Filter detuning is evident when filters are *operating* during extreme temperature changes. During *shipping*, temperature changes do not affect them. Once the filters return to room temperature, they should operate as specified.

Summary

There are many types of duplexers, cavities and isolators that can be used for transmitter and receiver isolation. The methods described in these five articles will assist with retuning cavities to provide the desired function with the antenna system.

(This concludes the article series.)

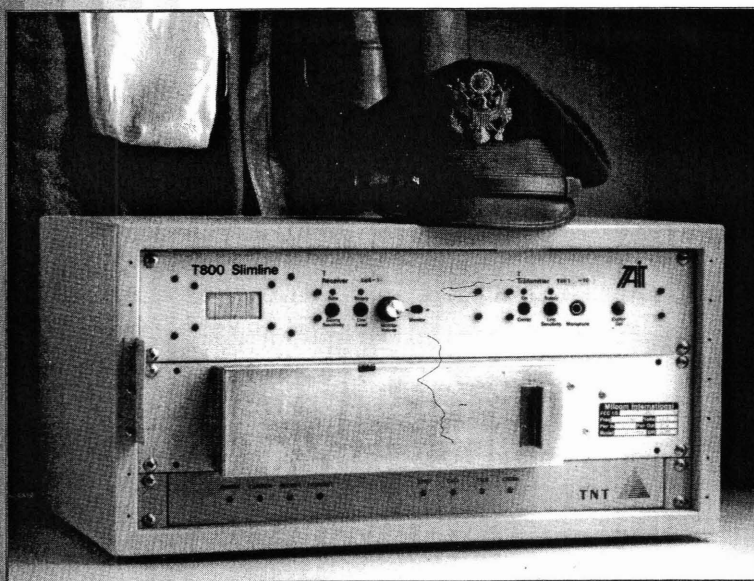
Duplexer series

Previous installments in this article series include:

- "How to use duplexers: Isolation Requirements" in the July 1994 issue.
- "How to use duplexers: The various types" in the August 1994 issue.
- "How to use duplexers: Installing and tuning" in the September 1994 issue.
- "How to use duplexers: Fine-tuning adjustments" in the October 1994 issue.
- "How to use duplexers: Combiners, multicouplers" in the November 1994 issue.

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Bridging devices in disparate bands

Interoperability often depends on linking networks operating outside of 800 MHz

Despite the steady migration of public-safety radio organizations to 800 MHz trunking systems over the past few years, many public-safety radio systems still operate in the VHF and UHF bands. Fortunately, the national frequency plan developed by the National Public Safety Planning Advisory Committee and adopted by the Federal Communications Commission in 1987 calls for certain 800 MHz frequencies to be set aside for the purpose of interoperability—or “mutual aid”—between public-safety entities operating in various frequency bands.

To review the requirements and objectives, I have excerpted a portion of Appendix C of NPSPAC's Channel Planning Analysis, as follows in the next section. (Visit www.pswn.gov/admin/librarydocs/800appendixc.pdf for the full document.)

The national plan requires that regional plans contain several sections concerning the implementation of the five nationwide mutual-aid channels. These sections typically define the use of each of these channels and provide detailed standards and operating procedures to govern the use of these channels.

Common Channel Implementation—The national plan sets forth the guidelines for using and implementing the five National Common Channels (see Figure 1). Four of these channels are dedicated as National Tactical Channels and one channel is dedicated as the National Calling Channel. The National Common Channels are to be available for use throughout the region.

The National Calling Channel, channel 601, is to be implemented as a full mobile relay, with wide area coverage transmitters to maximize coverage.

Figure 1: NATIONAL COMMON CHANNELS

Channel Name	Channel Number	Frequency (Mobile/base in MHz)
Call	601	821.0125/866.0125
TAC 1	639	821.5125/866.5125
TAC 2	677	822.0125/867.0125
TAC 3	715	822.5125/867.5125
TAC 4	753	823.0125/868.0125

Large system users (five or more channels) are required to monitor this channel and could be required to provide satellite receiver feeds into the wide coverage area.

The four National Tactical Channels are to be assigned throughout the region for use by all eligible entities. Large system users could be required to sponsor one or two localized mobile relays to cover specific geographic areas. The users of these channels must be eligible for licensing on other 800 MHz public-safety channels, but no special licensing is required.

Operation on Common Channels—The five common channels are only to be used for activities requiring intersystem communications between entities not already sharing communications systems and are not to be used for any daily operations. In emergencies, the channels may be assigned by the primary public-safety agency in that area. On all common channels, plain English and familiar words and phrases should be used. The calling channel is used to establish contact with other users and determine which tactical channel to

use. It is not to be used as an ongoing working channel. Tactical channels are reserved for interagency communications and are used as directed by the primary public-safety agency in the area. Tactical channels can be assigned by the various public-safety services, or they can be assigned by county or area.

Network Operating Method—A wide area network will be established on the National Calling Channel. The tactical channel communications systems will be implemented by volunteer entities, and each primary geographic section of the region is covered by at least one tactical channel.

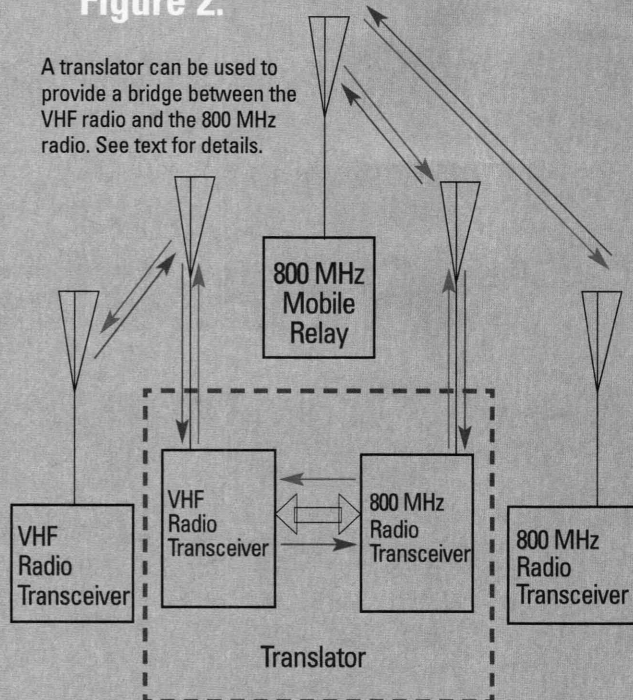
Coded Squelch on Mutual-Aid Channels—The National Common Tone Squelch of 156.7 Hz will be used on all equipment operating on the five common channels.

If these frequencies are used nationally, as the plan requires, then interoperability comes down to providing a link or bridge from public-safety radio systems operating in bands other than 800 MHz. In Figure 2, a conventional VHF radio is intercon-

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Figure 2:

A translator can be used to provide a bridge between the VHF radio and the 800 MHz radio. See text for details.



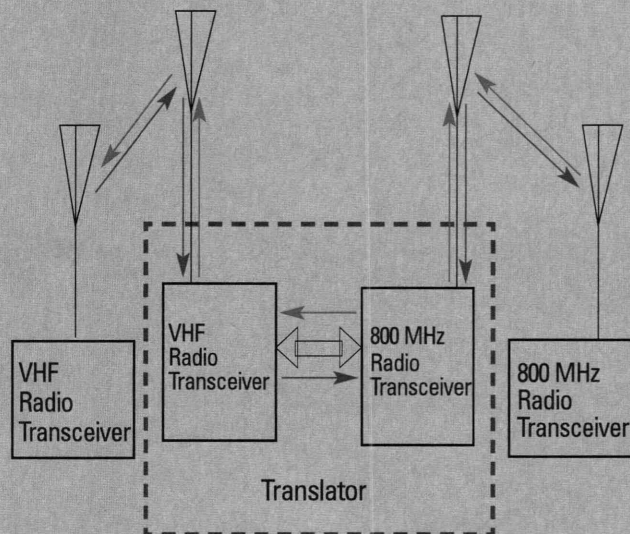
nected to a conventional 800 MHz radio to form a translator. The function of the translator is to receive radio transmissions on the 800 MHz system and retransmit (translate) them over the VHF radio system. The translator works bidirectionally—that is, it also picks up radio transmissions on the VHF radio system and retransmits them through the 800 MHz radio.

As shown in Figure 2, the audio out-

put of the VHF transceiver in the translator is fed to the exciter input of the 800 MHz transceiver. The control circuitry keys the 800 MHz transceiver, thus transmitting the signal to the 800 MHz mobile relay station. The 800 MHz mobile relay station then retransmits the 800 MHz signal to the 800 MHz mobile (or portable) radio transceiver. The signal follows the path of the blue arrows. When the 800 MHz mo-

Figure 3:

In this arrangement there is no 800 MHz mobile relay. The translator is the only bridge between the 800 MHz radio and the VHF radio.



bile transceiver is transmitting, the reverse operation takes place, and the signal follows the path of the red arrows.

Figure 3 shows another setup in which the translator is the only bridge between the 800 MHz radio and the VHF radio. The 800 MHz mobile relay station is omitted from the loop. It is important to note that the translator setups described above will only work on conventional 800 MHz frequencies.

More elaborate methods must be employed to link a conventional system at other bands to an 800 MHz trunking system. One solution can be found in the SmartBridge trunked radio interconnect systems from Melbourne, Australia-based Radio Systems Technologies. For example, the SB100 TalkGroup extender—designed for the company's SmartNet/SmartZone trunked radio networks—can be installed at any conventional site within radio range of a trunked site.

When the bridge is operating in the connect state, any radio operating within the repeater's radio range becomes an extension of the TalkGroup. Thus, the conventional radio effectively behaves as though it were a trunked

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radio itself. The connect on/off state is controlled by using "over-the-air" dual-tone multi-frequency signaling. A unique "trunked system busy" protocol ensures that users of conventional radios are immediately informed of unsuccessful attempts by audible alerts. (For more information visit www.rstradio.com.)

Another device that has gained popularity is the TRP-1000 transportable interconnect system from JPS Communications, a division of Raytheon. The TRP-1000 provides interoperable communications between organizations using different radios on disparate frequencies. It is a self-contained, portable unit consisting of a power source and multiple radios pre-wired to the company's ACU-1000 solution. The ACU-1000 is designed to provide on-scene interoperability by linking radios, SATCOM, cellular, Wi-Fi and landline phones directly and over IP networks.

The TRP-1000, integrated with the ACU-1000, can serve as a bridge between conventional systems as well as

Many public-safety agencies don't want to move to 800 MHz trunking systems.

between conventional systems and trunked systems. (More information on the TRP-1000 and ACU-1000 can be found at www.jps.com.)

Although trunked radio systems (800 MHz and other bands) currently serve a large number of public-safety agencies, many agencies continue to

operate conventional radio systems in the VHF and UHF bands. Of these, many—if not most—are pleased with their radio systems and don't want to move to 800 MHz trunking systems. Consequently, until the time comes when all public-safety agencies are on the same radio system, bridging devices such as those described in this space will be required to achieve interoperability.

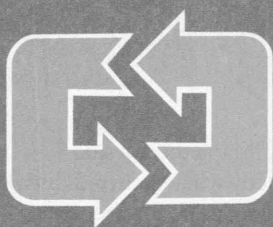
Until next time—stay tuned! ■

Editor's note: The following references were omitted from last month's Technically Speaking entitled "How to give radio signals a boost":

"Repeater Amplifier Systems: Principles and Applications," Ernesto Alcivar, ©1994 by TX/RX Systems, Inc. www.txrx.com.

"In Building Tricks," Dennis J. Burt, © 1994 by Intertec Publishing Corp.

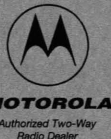
"Introduction to Signal Boosters," Jack Daniel, © 1996-2001 by the Jack Daniel Co. www.rfsolutions.com.



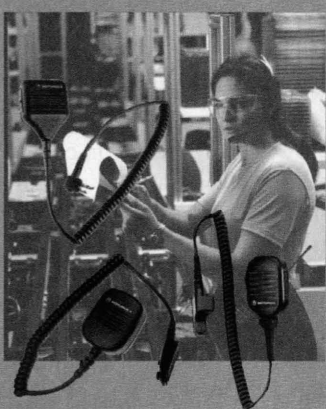
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Preventing microwave interference

Wind and other environmental influences play a major role in signal distortion

Over the past decade microwave links have proved to be a popular solution for the telecommunications industry. The relative ease and economy of their installation has fueled their deployment in an increasing number of point-to-point and point-to-multipoint applications—from communications backbones to branch links and distribution networks—in addition to applications in the broadcast industry and in private enterprises. With the rise of new cellular operators and new technologies, overall microwave network density is undeniably escalating.

However, the escalation of microwave communications has created new challenges. The greater the number of point-to-point links in a given area, the greater the potential they will interfere with one another. Because any distortion of the signal reduces the quality of service, controlling interference is now the mandate of any radio network operator and national regulatory authority. A good starting point for eliminating, or at least reducing, interference is the design and location of the signal source—the microwave antenna.

Figure 1 shows a typical radiation pattern of a microwave antenna. This example shows the main beam at 0 degrees, plus side lobes that are significant to about ± 90 degrees from the main beam. The side lobes can cause interference with adjacent point-to-point links and consequently must be minimized through careful antenna design and installation.

Radomes are used for two primary applications in radio-link antenna design. The first is to cover the antenna-feed system to protect it from the elements. In addition, a radome significantly reduces the wind load of

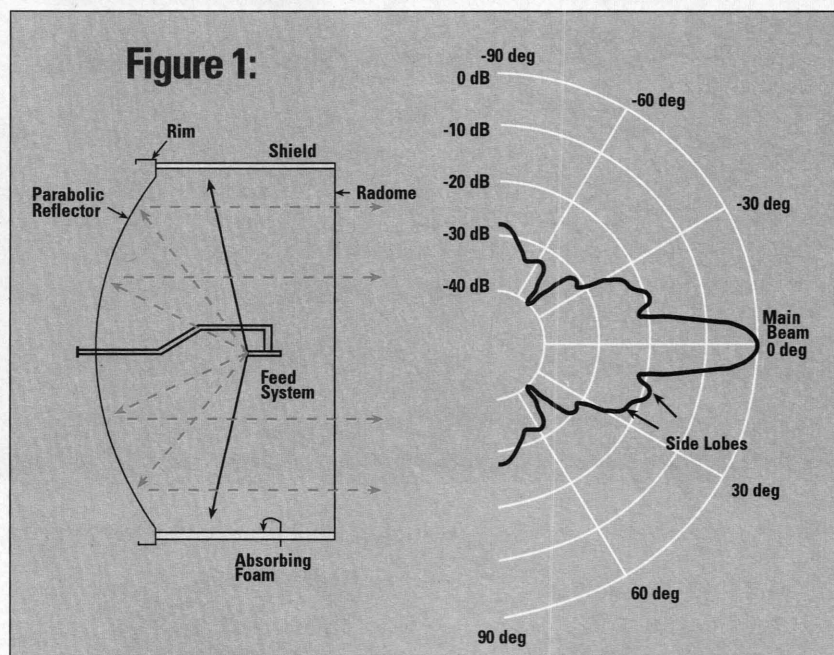
an antenna system by preventing the dish and shield from “catching” wind. However, both radome material and thickness need to be carefully considered to optimize the power transmitted through the radome, while at the same time ensuring the side lobes are not increased to detrimental effect.

Figure 2 shows a plot of the reflection characteristics at a plane-wall radome for different materials. Each of these materials are characterized by a relative dielectric constant of $\epsilon_r = 2$; however each material has a different loss parameter, $\tan \delta$, where $\tan \delta$ ranges from 0.0018 (low loss) to 18 (high loss). Figure 2 shows that for low-loss materials, there exists two distinct minimum values of the reflection coefficient, for which a radome wall will allow maximum transmission of incident power. These correspond to design values where the ratio of

radome wall thickness (d) to microwave wavelength in the sheet (λ) is close to either zero or 0.5.

The first case of $d/\lambda \approx 0$ is practically realized as $d < \lambda/10$ and leads to flexible radome materials with typical thicknesses of 0.4 to 0.6 millimeters—essentially as thin as is practical. Flexible radomes are commonly used for larger antennas (greater than 4 feet), to avoid the bulk and weight of solid radomes.

The second design case of $d/\lambda \approx 0.5$ is more complex, and leads to the design of solid radomes, which are more economical to produce at the smaller sizes (less than 6 feet). The practical implication of $d \approx \lambda/2$ is that the thickness of solid radomes is always dependent on the wavelength (and hence frequency) to be used in the application. Assuming a dielectric constant between 2.5 and 3, typical solid radome thicknesses for selected frequencies would be as follows: 6 mm (14 GHz), 4



mm (22 GHz), and 2.4 mm (38 GHz).

It is important to note that if a radome of incorrect thickness is used, the transmitted power will be reduced, and consequently the antenna gain also reduced. Greater radio power would then be required to achieve the desired radiation power, resulting in a corresponding increase of side lobe radiations. This illustrates the fact that correct radome design is critical not only

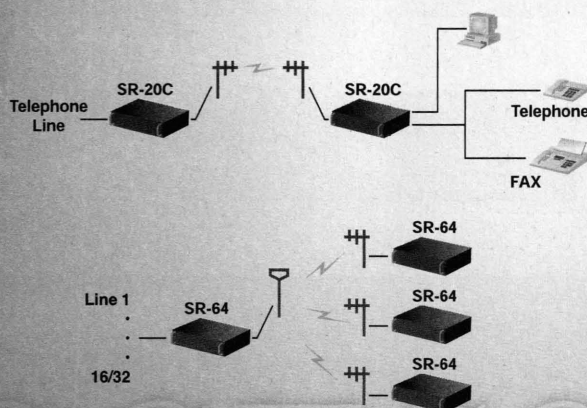
for optimizing link budget, but also for interference control, since an increase in side lobe radiations would raise the potential for interference.

Figure 2 is valid for the "ideal" case, where wave fronts hit the wall perpendicularly. Now consider the situation for signals not having this ideal orientation. Given the longer effective wave path through the radome material

as they hit the wall obliquely, the optimum thickness is now also dependent on the angle of incidence (θ), measured as the deviation from normal.

In practice, however, angles of incidence of up to 20 degrees have negligible effect on the optimum radome thickness. This is illustrated for flexible radome materials in Figure 3, which shows the relationship between angle of incidence (θ) and d/λ_0 (where λ_0 is the free space wavelength) for achieving 95% power transmission through materials with different dielectric constants. For values of θ up to 20 degrees, the optimum thickness is barely impacted—particularly for low-loss materials, which should be those considered for radome design purposes.

A similar relationship holds for solid radomes. A small degree of tilt in the main beam—about 5 degrees—actually improves the performance of the

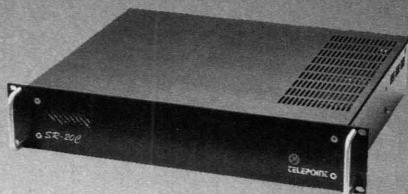


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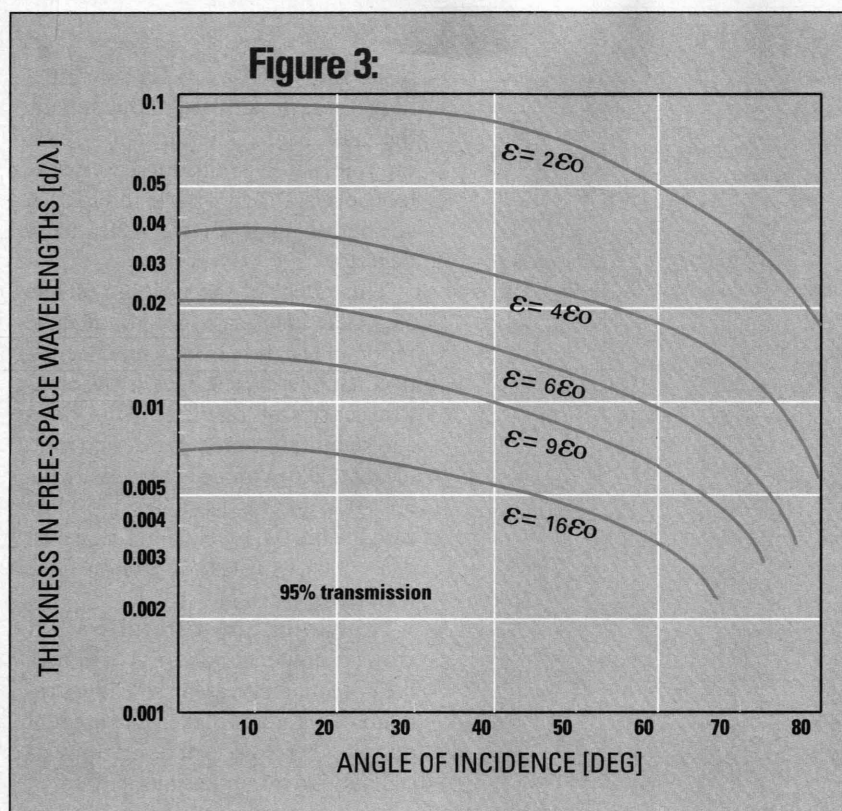
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**A small degree of tilt
in the main beam—
about 5 degrees—
actually improves the
performance of the
antenna.**

antenna, by directing spurious reflections within the antenna away from the microwave feed system.

The influence of typical thin-wall or flexible radomes can be seen in Figure 4, which compares the radiation patterns—with and without radomes—of microwave antennas operating at 6.4 GHz and 33.4 GHz. At 6.4 GHz it is evident the radome has negligible effect on the radiation pattern. However, at 33.4 GHz the gain of the antenna is decreased by 1 dB due to attenuation by the radome. Once again, to achieve the same link budget, the radio power has



The standard-performance microwave antenna consists of an open dish and a feed system. Usually lacking a radome (although a molded radome is an option), standard-performance antennas are economical solutions for specific applications. Aside from the lack of environmental protection of the feed system, the main drawback is the diffraction of microwave power at the rim of the dish, resulting in significant backward reflections at azimuth angles of ± 100 degrees that can seriously interfere with adjacent point-to-point links.

To literally block these backward rim reflections, antenna designers place a shield around the circumference of the antenna, to which a planar radome is usually attached. These "high-performance" microwave antennas may be further enhanced by the application of an absorbing foam to the inside of the shield, resulting in "ultra-high-performance" microwave antennas. The foam absorbs spurious reflections within the antenna and dramatically improves performance by limiting the side lobes.

Radiation-pattern envelopes for standard-, high-, and ultra-high-performance antennas are compared in Figure 5. The improvement in side lobe-reflection control of the ultra-high-performance antenna over both other antennas is evident.

Interestingly, though, the high-performance antenna exhibits poorer performance than the standard-performance antenna between 20 and 60 degrees—the result of additional re-

Selection of the appropriate microwave antenna depends on the intended application and the expected interface potential in a given area.

flections off the shield. It nevertheless proves significantly better at preventing backward reflections. Selection of the appropriate microwave antenna clearly depends on the intended application and the expected interference potential in a given area.

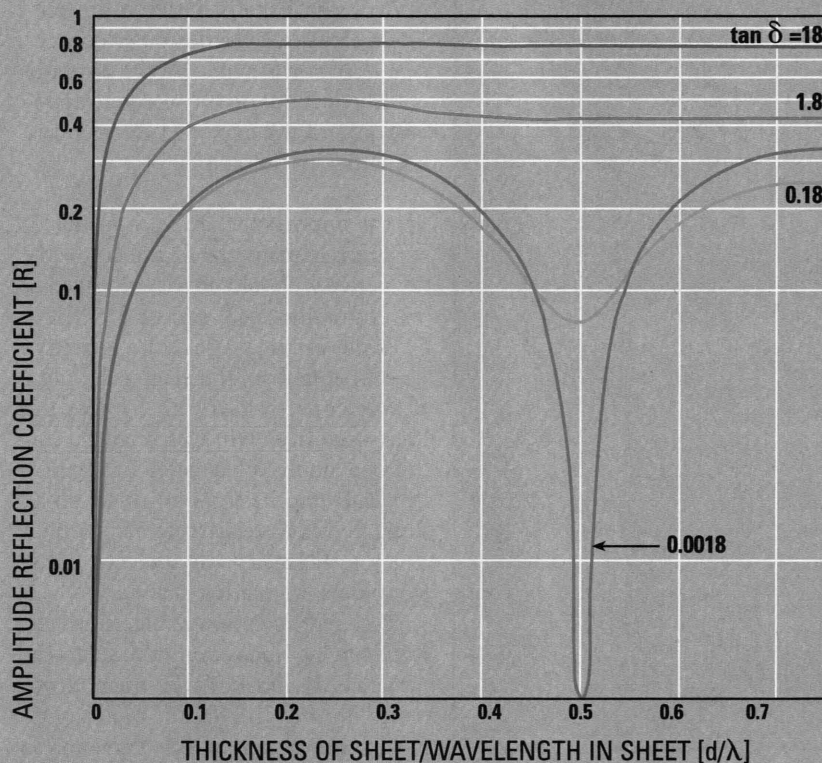
It is important that, once installed, the performance of a microwave network should not deteriorate due to environmental impact. While a radome might protect the sensitive feed system from the elements, only a stable construction can protect the dish itself from wind. Mechanical stability of an installed antenna is critical for it to maintain its point-to-point link, as well as restricting its potential for interference with adjacent links, should its orientation change.

Different antenna manufacturers use different methods of rating antenna resistance to wind. Radio Frequency Systems defines the operational wind-speed rating of an installed antenna as that for which temporary deflection of the main beam is within one-third of the half-power beam width of the antenna. (Half-power beam width is defined as the angle, relative to the main beam axis, between the two directions at which the measured co-polar pattern is 3 dB below the value on the main beam axis.) Within this operational wind speed—of which typical values are 120 to 140 miles per hour—the point-to-point link will be satisfactorily maintained.

Other standards consider the operational wind speed as that for which the main beam is not deviated by more than 0.1 degree. Whatever the method used, it is important to take the deflection of the mounting structure into consideration during calculation of the beam deflection.

The positional mounting of antennas must also be considered by operators seeking to minimize interference. Typical multi-antenna tower installations, with their many side lobe radiations, are a breeding ground for interference. In other cases, the mounting structure may directly impact the performance of the microwave link.

Figure 2:



to be increased by 1 dB, causing a higher interference potential outside the main beam. The presence of the radome also leads to increased side lobe levels, clearly visible in Figure 4 at azimuth angles between 20 and 60 degrees.

This effect of the radome on the 33.4 GHz antenna is because flexible radome design becomes more sensitive, at higher frequencies, to the practical constraints on material thickness and stability. Whereas the design ratio of $d/\lambda \approx 0.01$ can be achieved for the 6.4 GHz antenna, the best possible case for the 33.4 GHz antenna is just $d/\lambda \approx 0.05$, which is not as close to the ideal zero.

No matter how carefully a microwave antenna radome is designed, the potential increase in side lobes remains. This must be taken into account during other aspects of design and installation to minimize interference.

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through shielding and the generation of reflections from outside of the antenna itself; this is particularly true when antennas are mounted on the face of buildings and solid towers. Significant reflections could arise when an antenna is mounted too close to a solid structure to the side. Such structure-generated reflections are likely sources of interference, and often are not taken into consideration by operators during installation.

Managing interference is not a new issue, but it has certainly become more critical as the prevalence of radio networks has grown. Judicious antenna selection, design and installation is more essential than ever to minimize the interfering effect of spurious emissions, and to maximize performance of the microwave link network. ■

Daniel Wojtkowiak is global product manager, vice president-microwave antenna systems for Radio Frequency Systems. He holds a doctorate in engineering based on his studies of multi-frequency radar. He joined RFS kabelmetal in 1989, and in 1993 became head of the engineering group responsible for the development of radio-link antenna and passive waveguide components. In 1999 he became product manager-microwave antenna systems, responsible for Radio Frequency Systems' European region, and in 2003 was appointed to his current position.

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Figure 4:

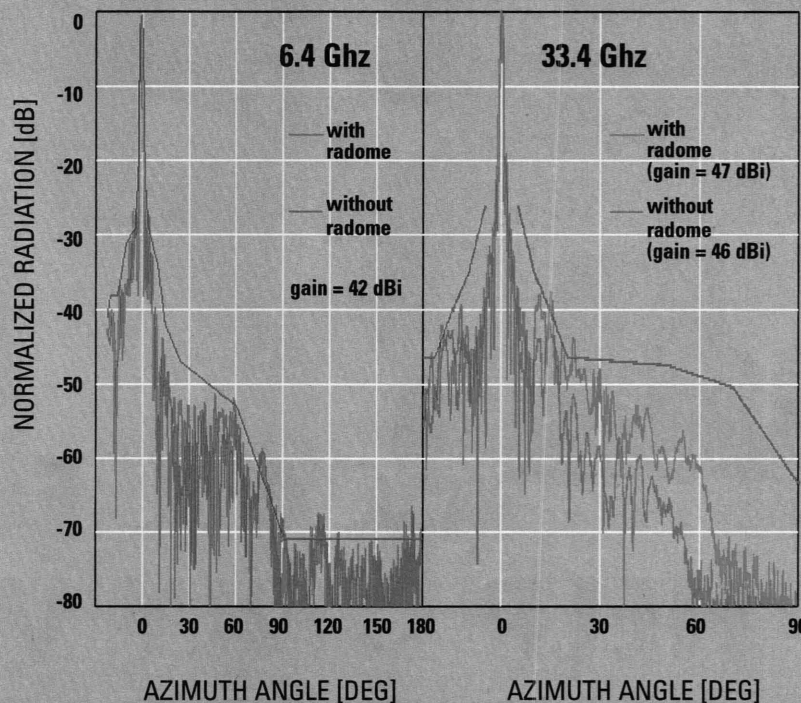
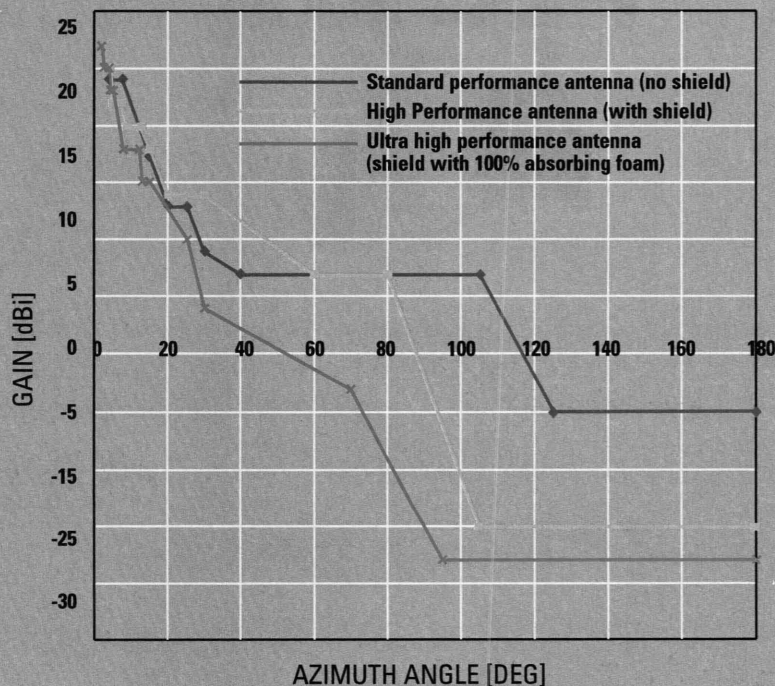


Figure 5:



Taking control of Project 25 systems

Test and evaluation of control channels is the key to solving the interoperability puzzle

Protocol is a crucial factor in achieving interoperability between mobile and portable digital radios operating on APCO Project 25-compliant networks. Equally important are test solutions that effectively log and analyze control messages to determine system compatibility.

The traffic and control channel fea-

tures for both conventional and trunked P25 systems are not specific to any one P25 manufacturer or P25 radio frequency subsystem (RFSS) configuration. All mobiles supporting P25 formats should be able to access any P25 network as specified in the TIA/EIA-102 standard, which defines the operation of P25 systems. As radio systems

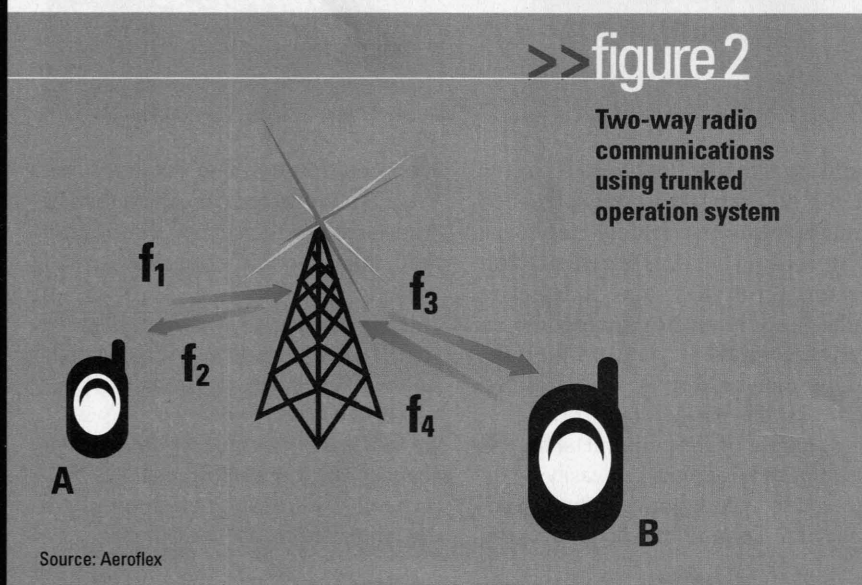
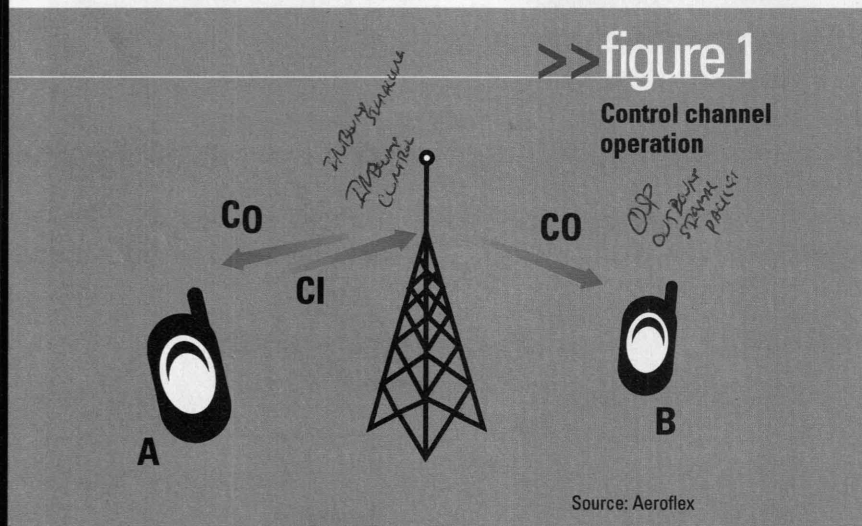
continue to move from analog to digital, an understanding of how the control channels operate will help communications technicians ensure interoperability with new P25 systems.

In the P25 standard, control channel operation maintains compatibility between conventional operations in that the modulation format and the bit rate are the same. In the P25 standard, the control channel and the traffic channel is a 9600 b/s channel using C4FM modulation to impart the digital 1s and 0s to the receiving radio. Trunked operations use packet-access techniques and require a registration process to coordinate user access, while the conventional version permits users to control their own access. Figure 1 provides an example of how a control channel operates.

As illustrated, the repeater is transmitting over an outbound control channel (CO) that both mobiles are monitoring. The outbound control channel sends information in "packets" called outbound signaling packets (OSPs).

Assume that mobile A wants to place a call to another user on the system. It transmits a request to the repeater on an inbound control channel (CI) using inbound signaling packets (ISPs). The repeater station and network process the request and then send an alert on the outbound control channel that the designated user, mobile B, recognizes as part of its group or as a signal designated specifically for itself.

As shown in Figure 2, the calls are then assigned to particular traffic channels that operate on separate frequencies (f_1 , f_2 , f_3 , f_4). The transmit and receive frequencies for the mobiles are offset from each other (f_1 and f_2) (f_3 and f_4) using a duplex offset. In the 800 MHz band, the offset is 45 MHz.



Control of these resources is accomplished through the control channel that the mobile unit monitors continuously when not engaged in a communication session with another user or with the network. When it is engaged in communication, link control octet (LCO) messages ride along with the voice data on the traffic channel to control the communications link.

The ability to capture and save over-the-air control channel messages provides unique testing capabilities in P25 systems, including:

- **Verification that specific messages are being generated correctly from the P25 RFSS infrastructure:** System operators need to be 100% certain that the messaging is exact and is

By providing the ability to log a control channel, design engineers and technicians can capture a suspect control channel in one location.

being sent at the right time and in sequence. Improper timing or corrupt messages can effectively take an entire group of radios off the network.

- **Verification of encryption:** The P25 standard allows for the traffic chan-

nels to be encrypted. Logging the control channel can show the system operator whether the proper encryption algorithm identifier for either AES or DES encryption and the proper key identifier is being sent.

- **Field troubleshooting of radio interoperability problems:** Since the P25 standard is an open standard, there are many potential suppliers of P25 equipment. Although the standards are well-written and provide most of the information about implementing the proper protocol, sometimes differing interpretations of these standards lead to radios that simply will not work on a given system. The ability to verify the control channel and monitor how the radio responds allows the system engineer to verify radio interoperability with a particular control channel.

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By providing the ability to log a control channel, design engineers and technicians can capture a suspect control channel in one location. It can then be analyzed in a lab environment to see how radios are responding to some of the message elements and to compare various control channel implementations.

The P25 control channel messages are provided in two formats: single- and multi-block trunking signaling blocks (TSBKs). Most of the control channel

information is sent via the single-block TSBK, with the multi-block TSBKs reserved for messaging that requires more extensive information. Figure 3 defines the single-block TSBK formats.

There are two multi-block formats. One uses the normal unconfirmed data header (UDH) format, with a special service access point (SAP) identifier in the header that causes the message to be processed as multi-block trunking information.

The other type is very similar, except

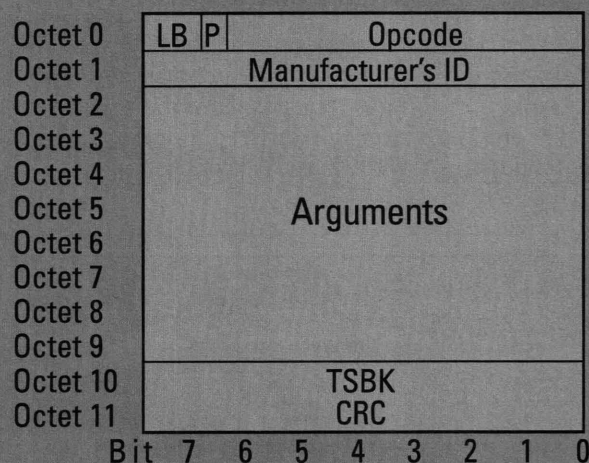
that the header contains additional information that includes embedded operations codes and trunking-specific parameters for use in the multi-block TSBK. System implementation and the type of messages sent will determine which of these message formats should be used. Figure 4 defines the multi-block TSBK formats.

The TIA/EIA-102 standard defines several control channel messages, both in the multi-block and single-block formats. These can be broken down into three primary messaging groups: voice service, data service and control-and-status OSPs.

Voice service messages deal with the transactions, or messaging, sent between the control channel and the user to establish voice communications. Data service messages act much in the same manner to support data calls. However, the majority of the control channel signaling occurs for control-and-status information, which is critical to both the subscriber (user) and the system infrastructure because it sets up communication parameters before voice or data communications can occur.

To gain a full knowledge of the system operation, it is desirable to log both

>>figure 3

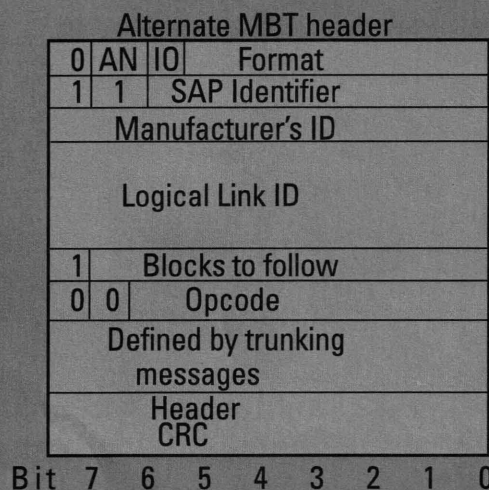
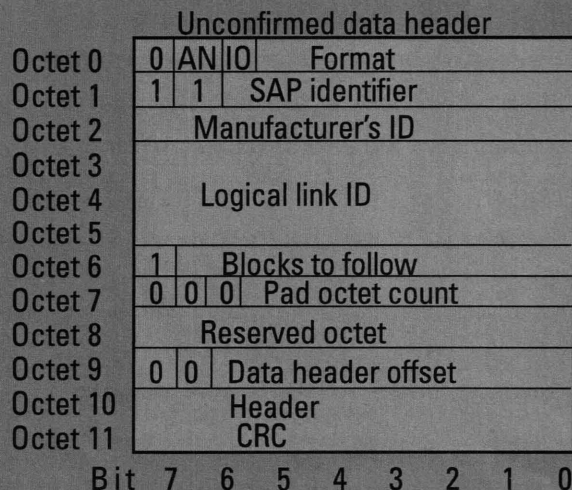


Single block packet structure format

Source: Aeroflex

>>figure 4

Multiple block TSBK formats



Source: Aeroflex

the OSP and ISP messages to compare how the mobiles and network operate together. Logging the LCO messages that accompany the voice or data information also is beneficial, particularly when the radios are used in a conventional repeater application.

A communications engineer can easily log control channel messages by utilizing a P25 test system (such as Aeroflex's IFR 2975). When the receiver has been tuned to the proper frequency, such a test system will "listen" for various messages coming through. The system matrix allows the user to determine what messages to log and at what level. The system also can log the raw C4FM symbols before any error correction, the octets after error correction or the actual messages decoded with the actual message elements.

Once the receiver is tuned to the proper frequency, the test system will automatically decode and log the mes-

sages based on how the matrix is set up. Since the unit has only one receiver, only one type of message can be de-

An advanced P25 test system helps engineers and technicians quickly ascertain problems associated with control channel implementation

coded at any one time. The system will then log the messages to a file on the unit's internal hard drive and will display the sequence of messages received with a time stamp.

As can be seen by the above steps in

utilizing an advanced P25 test system, engineers and technicians can quickly ascertain interoperability problems and other problems associated with control channel implementations to assist in the design and field test of P25 systems. ■

Rob Barden is senior product marketing manager for the radio test set line at Aeroflex's Wichita, Kan., headquarters. He is a 22-year veteran of the test and measurement industry and graduated from Devry Institute in 1983.

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Cognitive radios close to major advance

Software-assisted radios are the first step toward true software radios

Cognitive radio systems in general, and true software radio (TSR) products in particular, are on the verge of revolutionizing the communications industry. TSR systems, by providing all signal processing in the software, are approaching

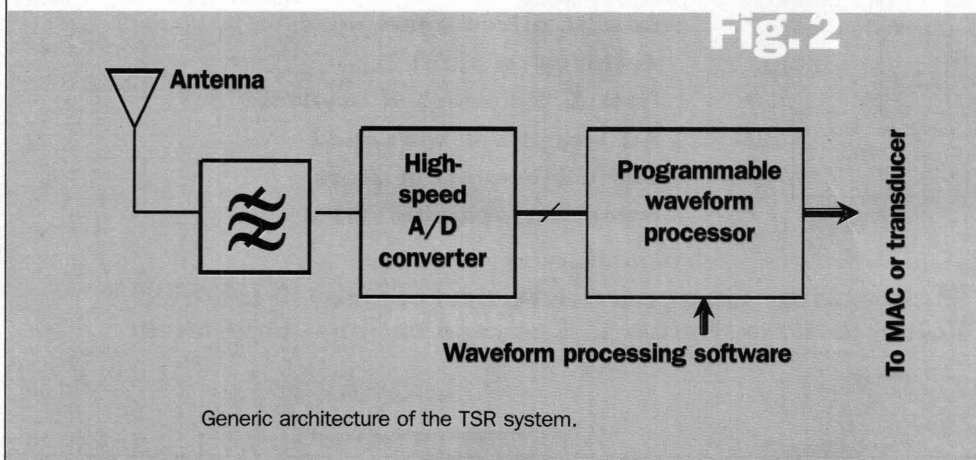
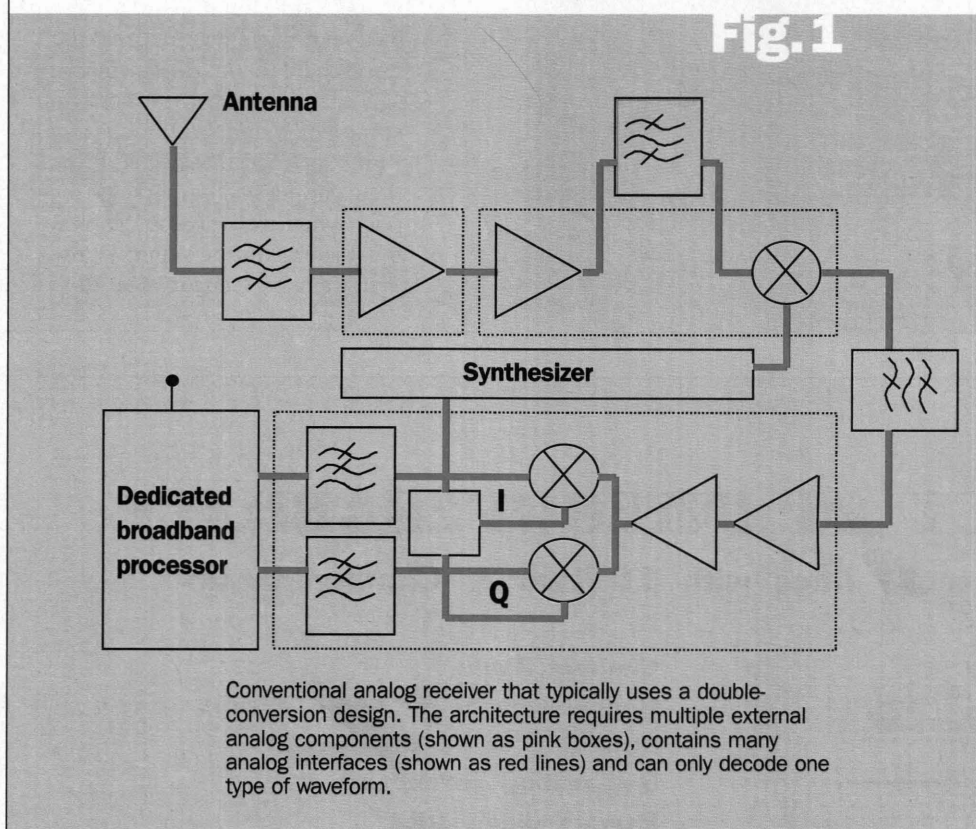
the ultimate software-defined radio (SDR) systems goal of seamless operation in highly fragmented, multi-terminal/multi-frequency communication environments.

Such advancements are achieved by converting the signal into digital form

immediately after receipt by an antenna and by dynamically downloading software corresponding to the specifics of the received signal. The progress toward SDR's ultimate goal of seamless operation in a multi-standard environment will be driven by the rate of advancement in microelectronics, particularly the increase in microprocessor computational power and a reduction in power dissipation.

The Software Defined Radio Forum—an industry initiative striving to accelerate development and deployment of SDR systems—stipulates that SDR products possess two fundamental features: flexibility toward operational standards and independence from carrier frequencies. While TSR systems possess both features, software-assisted radio (SAR) systems, in which only a portion of signal processing is done in the software, lack the reconfigurability—which must include both carrier frequency and communication protocol—to ultimately provide both features. SAR systems (Figure 1) thus can be viewed only as an intermediate step on the evolutionary path toward TSR.

In a TSR, the analog-to-digital converters (ADCs) are placed as close as possible to the antenna (Figure 2), which places great demands on the ADCs' performance. All subsequent signal processing of the digitized antenna output is done by fast logic circuits and fast microprocessors



using downloadable signal-processing software selected according to a system's operational environment.

Effective quantization of the radio signal at the antenna enables fast re-configuration of the air interface parameters of the communication terminals. The dynamic switching of frequencies and communication protocols in the user's terminals enable the

come to life in the not-so-remote future.

For now, however, the more realistic goal (as far as handsets are concerned) centers on developing TSR systems capable of implementing a couple of protocols per unit and providing a steady supply of power sufficient to perform

signal and data processing. These criteria limit the range of current possible full TSR implementations to a stationary environment such as base stations or vehicle-based systems. Next-generation TSR components will implement digital signal process-

Architecturally, TSR is best defined as the software implementation of the radio transceiver receiving digitized down-converted signals from an antenna.

remote reconfiguration of the terminal by adding or removing system software components, with the result being greater flexibility.

Thus, TSR is the only technology that currently shows promise in delivering the ultimate SDR goal of a truly "universal" radio terminal. Other approaches are vulnerable to changes in applied standards and to introduction of new functionality.

While the significant lack of sufficient processing power currently prevents SDR from yet becoming a full-scale reality in low-power applications such as handsets, the unabated doubling of processing power each 18 months (or Moore's Law) allows some assurance that SDR-based terminals and network nodes will

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Table 1**Estimated performance characteristics of a delta-sigma converter**

Converter order	Second	Third
SNR[dB] for $\Delta F = 30$ MHz	55	80
SNR[dB] for $\Delta F = 20$ MHz	65	90
SNR[dB] for $\Delta F = 10$ MHz	75	100
SNR[dB] for $\Delta F = 2$ MHz	100	

ing (DSP) primitives directly into the silicon, allowing for implementation in lower-power consumer devices.

An SDR suitable for commercial narrowband and broadband applications will typically cover the frequency spectrum between 400 MHz and 6 GHz. This range embraces most of the existing and emerging standards along with likely future developments.

The basic ingredients in the design of TSR hardware are ADCs, as well as reconfigurable hardware such as field programmable gate arrays, programmable logic devices, DSP boards and general-purpose computers. The embedded software can reside in all the programmable entities used in the design.

The TSR architecture must be able to accommodate operation in different environments characterized by different standards, carrier frequencies, power levels and bandwidths. Architecturally, TSR is best defined as the software implementation of the radio transceiver receiving digitized down-converted signals from an antenna. Digitization of the wireless signals' function at the antenna dramatically simplifies the implementation of transmitters and receivers. In addition to using direct down-conversion receivers versus traditional superheterodyne receivers, the architecture requires few external analog components and can be programmed to process any type of signal or multiple types of signals.

TechnoConcepts chose the Linux operating system for its TSR implementation because Linux guarantees maximum accessibility to all computer resources—such as device drivers for input/output (I/O) operations—and because it is in widespread use. Delta-sigma data conversion circuits capable of operating at clock rates in excess

of 5 GHz power the transceiver.

Delta-sigma converters digitize signals by modulating the analog input into a high-speed one-bit digital data stream that is subsequently processed digitally to produce a high-resolution word stream at a slower data rate. The converter is a closed-loop system in which the order of the loop and the input bandwidth may be traded for resolution. A plot of the ideal resolution for a given relative bandwidth and loop order is shown in Figure 3.

One way to improve upon the delta-sigma data converter technology is to create an architecture that simultaneously extracts the modulation from an incoming wireless signal and digitizes it with extremely high resolution. A dynamic range of 55 db to 100 dB (depending on bandwidth) is achievable using this architecture. The performance analysis of high-order delta-sigma ADC converters operating at 5 GHz shows that ex-

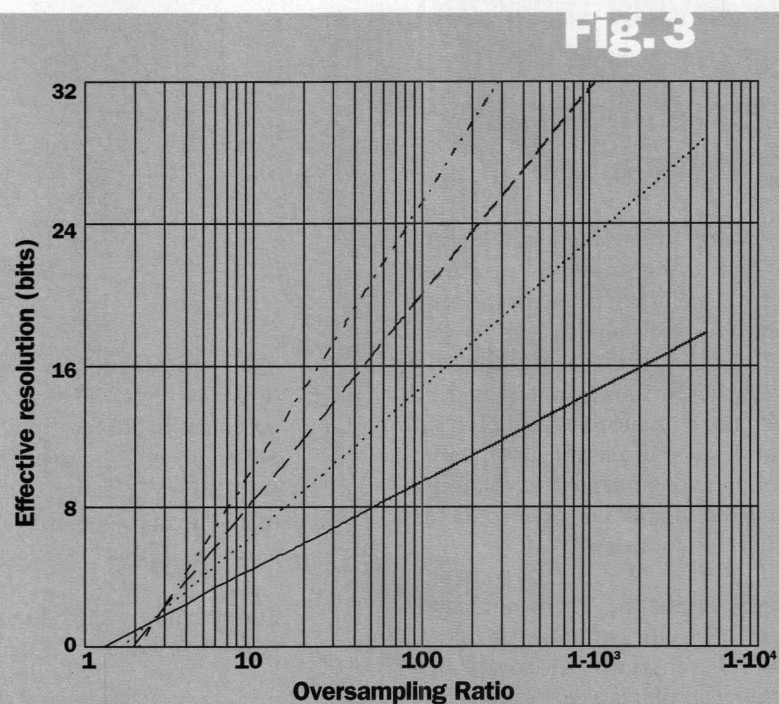
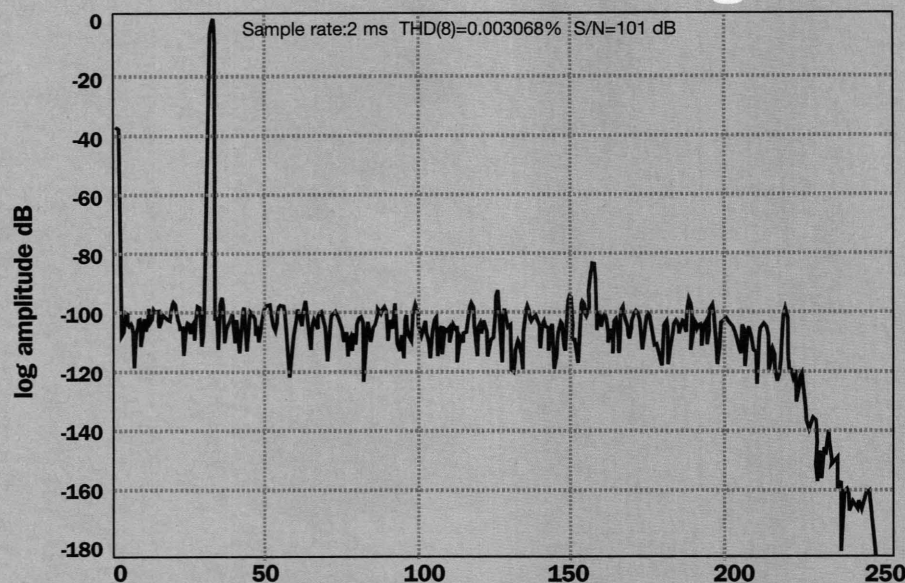


Fig. 3
A plot of the effective resolution of a delta-sigma converter for the first, second, third and fourth order modulators using one-bit quantization. The x-axis represents the ratio of clock rate to maximum input bandwidth (oversampling ratio). The y-axis represents resolution in bits.

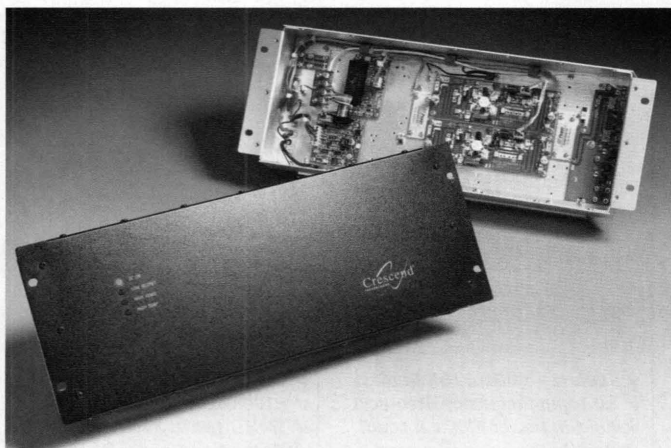
Fig. 4



Performance of CMOS technology converter at a clock rate of 10 MHz.

pected signal-to-noise (SNR) ratio depends on the operational spectrum width ΔF and the order of delta converters. The estimated SNR values in decibels for different converter orders as well as operational bandwidths, measured in MHz, are shown in Table 1 on page 26.

TechnoConcepts already has demonstrated 100 dB dynamic range using CMOS (complimentary metal oxide semiconductor) technology at a clock rate of 10 MHz, and recently demonstrated roughly a 50 dB dynamic range in its preliminary demonstration system using GaAs (gallium-arsenide) technology at a clock rate of 1.8 GHz. These results are shown in Figures 4 and 5.



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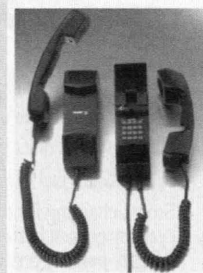
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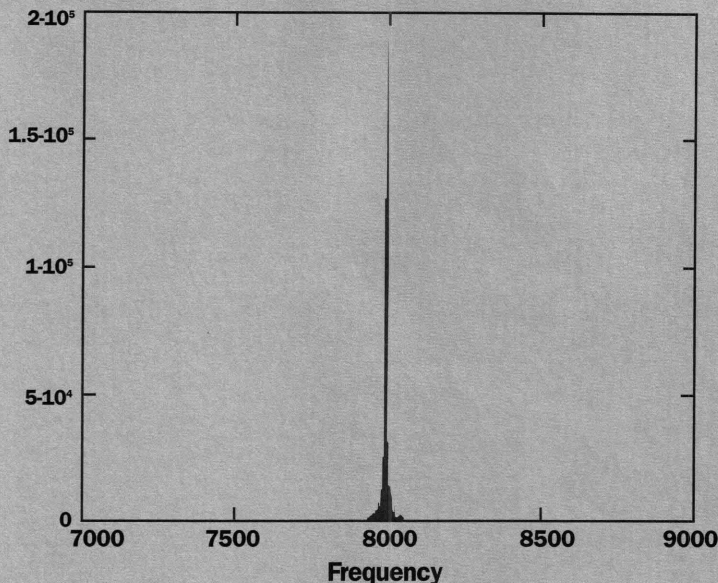
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As TSR products with more powerful processing capabilities evolve, a wider range of applications will emerge. For example, a single protocol cellular phone being connected to a variety of networks through a software radio base station that serves both as a repeater and (when necessary) as a protocol translator. In this particular application, software radio handsets are not required to achieve universal access, at least in the nationwide context.

A similar concept for automobiles can be illustrated. In this case, the car has a dual mode transceiver capable of transmission and reception using a proprietary protocol and also is capable of receiving navigation signals from the Global Positioning System. This capability permits the car to transmit position information to the base station

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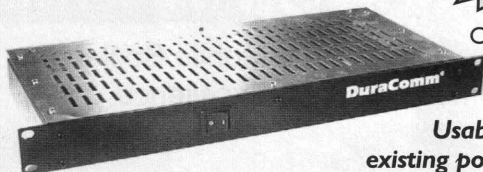


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and further allows it the universal access that is enjoyed by the handset.

The use of software radio technology would enable the car to access any cellular service and potentially be able

received signals. Two very different types of SDR products are emerging.

The first one—SAR—is a hybrid, where a portion of signal processing (usually at high or intermediate fre-

baseband immediately after receipt by an antenna. As a result, they satisfy both above-mentioned criteria. Such systems correspond to the spirit and expectations put on SDR.

Like any ultimate goal, it can be approached in an evolutionary manner, with each microelectronics advance bringing us toward multi-terminal/multi-frequency operation. ■

In bandwidth-intensive broadband applications software radio technology allows service providers to dynamically use unlicensed frequency domains.

to roam internationally. Furthermore, specialized services and capabilities (such as automobiles serving as repeaters for other automobiles and/or handsets) can be implemented without disrupting the vehicle's access to standard services. In this configuration (where both the base station networks and the access device utilize software radio technology), any wireless access device can communicate with any other wireless access device with multiple choices as to network services.

In such an environment, the user can select the service best able to meet his needs, based on bandwidth, cost or latency. This will create a competitive cognitive radio environment where limited bandwidth use can be efficiently allocated so that the user pays only for the bandwidth he needs and no usable bandwidth remains idle.

In broadband applications, that are bandwidth intensive, software radio technology allows service providers to dynamically utilize unlicensed frequency domains to meet additional client needs when saturation has been realized on a certain frequency domain. Residential customers could be receiving video feeds on unlicensed frequency domains while business customers concurrently receive video feeds on licensed frequency domains.

The ideal SDR products must possess two fundamental features—flexibility toward operational standards and independence from carrier frequencies of

quencies) is done in hardware and the rest in software. Such systems achieve partial independence from carrier frequencies, but are still dependent on analog components for filtering and other signal processing operations, making them reconfigurable but not frequency agile.

The second type—TSR—uses direct down-conversion from RF to

Ronald M. Hickling is TechnoConcepts' co-founder and chief technology officer. He holds a masters degree in electrical engineering from UCLA and has more than 20 years experience in developing communications systems for defense and commercial contractors, beginning his career with Hughes Space and Communications in 1980. Hickling founded TechnoConcepts in 1991 and has been awarded three patents and currently has two patents pending.

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Antenna phasing 101

Proper phasing is the key to achieving the desired radiation pattern

The method used to achieve antenna gain combines two or more antennas to produce an array. Redirecting or focusing the radiated electromagnetic energy in the desired direction achieves the antenna gain. Figure 1 shows a typical vertically oriented half-wave dipole and the re-

sulting radiation pattern. Notice that the radiation pattern resembles a doughnut. The radiation at the higher angles is wasted because the radio system is designed to talk to ground units.

In land mobile radio, the half-wave dipole is the standard reference antenna (Figure 1). That is, the gain of the

antenna is referenced to a half-wave dipole and stated in decibels referenced to a half-wave dipole (dBd). Suppose that two half-wave dipoles are stacked vertically as depicted in Figure 2. This arrangement will compress or squash the doughnut in the vertical plane and expand it horizontally, similar to compressing a round balloon. If this array produces field intensity at a given location that is 3 decibels greater than the half-wave dipole, it is said to have a gain of 3 dBd in that direction. Theoretically, for a truly omnidirectional an-

Figure 1:

Cross-section view of the radiation pattern of a vertically oriented half-wave dipole.

Cross-section view of the doughnut radiation pattern

Vertically oriented half-wave antenna

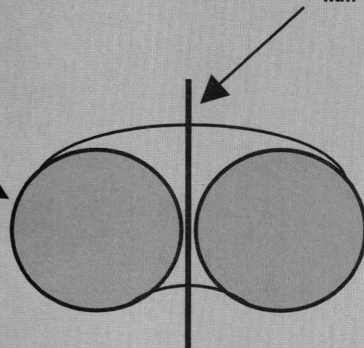
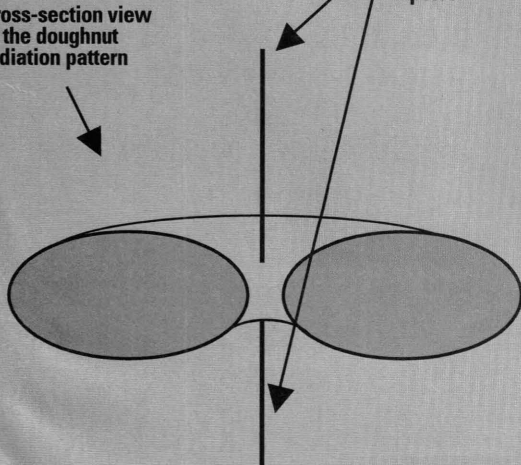


Figure 2:

Cross-section view of the doughnut radiation pattern

Two vertically stacked half-wave dipoles



Cross-section view of the radiation pattern of two vertically stacked half-wave dipoles. Notice that the doughnut radiation pattern has been compressed in the vertical plane and extended in the horizontal plane.

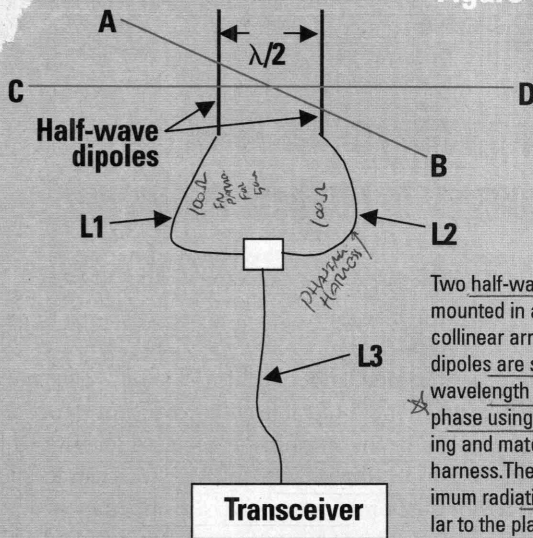
By directing most of the radiation parallel to the horizontal plane, antenna gain is achieved.

tenna, the field intensity will be equal (or nearly so) at all locations equidistant from the antenna in the horizontal plane. In the practical world, this will not be the case because there are different obstructions and path terrain in different directions.

Antennas stacked vertically as shown in Figure 2 are said to be vertically collinear. Thus, by directing most of the radiation parallel to the horizontal plane, antenna gain is achieved. An antenna array will produce stronger field intensity at a given distance from the antenna or will produce the same field intensity at a distance further from the antenna as compared to a single half-wave dipole.

Antennas can be combined in the

Figure 3:



Two half-wave dipoles are mounted in a horizontally collinear arrangement. The dipoles are spaced one-half wavelength apart and fed in phase using the proper phasing and matching harness. The direction of maximum radiation is perpendicular to the plane of the dipoles as shown by the green line AB. In the direction parallel to the plane of the dipoles, the radiation cancels, producing minimum radiation as shown by the red line CD. See text for more details.

horizontal plane, as well. Figure 3 shows how combining two half-wave dipoles in the horizontal plane affects the radiation pattern. If the dipoles were fed 180° out of phase, the pattern would shift by 90° such that the figure-eight pattern would lie in the dipole plane. Figure 4 shows the radiation pattern in the horizontal plane. The horizontal collinear mirror-mount antenna was popular with truckers using CB radios because this would project the signal in the direction of the highway path.

Proper phasing of the antennas is necessary to produce the proper match and radiation pattern desired. A phasing harness is necessary to provide the proper 50-ohm impedance to the transceiver and the proper phasing of the individual antenna elements. Figure 3 shows how two individual 50-ohm antennas are connected together such that the signal is fed to both antennas in phase and the point where all antennas are combined is 50 ohms. Note that each antenna depicted in Figure 3 is 50 ohms at the operating frequency. Because we are combining two antennas in parallel at the connection point, the impedance of the

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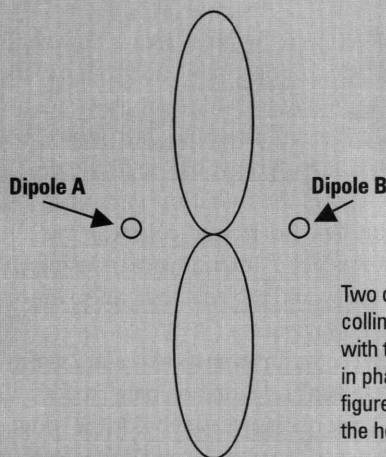
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Figure 4:



Two dipoles in a horizontal collinear arrangement and with the two dipoles fed in phase produces the figure-eight pattern in the horizontal plane.

Equation 1:

$$Z_0 = \sqrt{Z_1 \times Z_2}$$

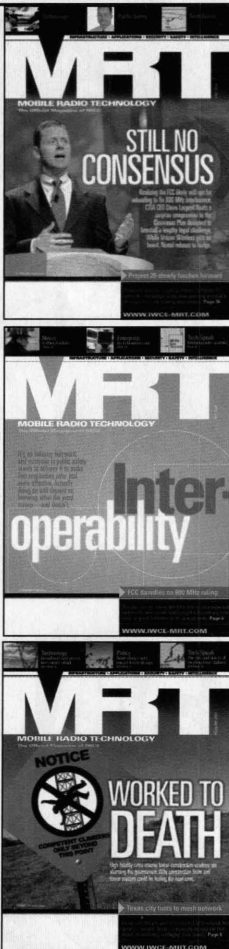
Where Z_0 is the characteristic impedance of the quarter-wave cable, Z_1 is the impedance at one end of the cable and Z_2 is the impedance at the other end.

connecting cables-L1 and L2-must be 100 ohms at the point of connection. Two 100-ohm impedances connected in parallel produces an impedance of 50 ohms.

Proper phasing of the antennas is necessary to produce the proper match and radiation pattern desired.

The characteristic impedance of cables L1 and L2 must be such that the proper length of cable will provide an impedance transformation from 50 ohms at the antenna end to 100 ohms at the connecting point. A quarter wavelength of cable that provides the proper characteristic impedance can transform the 50-ohm impedance of each antenna to 100 ohms at the connection point. The required characteristic impedance of cables L1 and L2 can be found from the formula in Equation 1. In this example, the calculation is shown as:

$$Z_0 = \sqrt{50 \times 100} = \sqrt{5000} = 70.7$$



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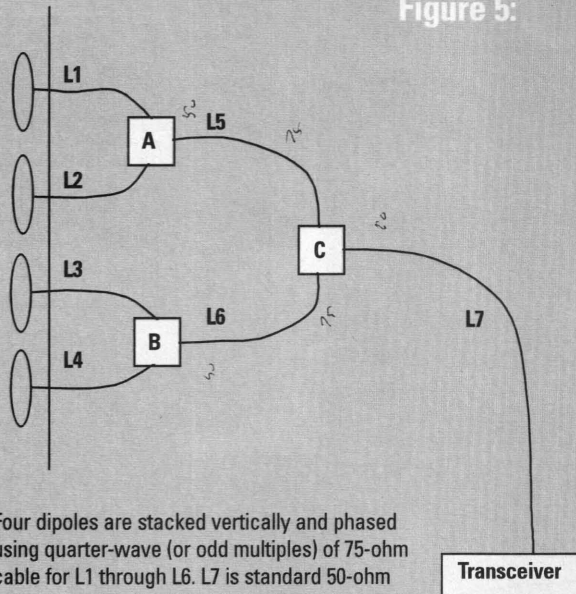
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Figure 5:



Four dipoles are stacked vertically and phased using quarter-wave (or odd multiples) of 75-ohm cable for L1 through L6. L7 is standard 50-ohm cable. Lengths of cables L1, L2, L3 and L4 must be identical. Similarly, lengths of cables L5 and L6 must be identical to feed the various dipole elements in phase. The length of cable L7 is not important as far as phasing the dipole elements.

Thus, the characteristic impedance of cables L1 and L2 must be 70.7 ohms. Because 75-ohm cable is readily available, it is generally used to make the phasing harness. Standard 50-ohm impedance cable is used for cable L3, which extends from the connection point to the transceiver.

Connecting four dipoles is done in a similar manner (Figure 5). Cables L1 and L2 connect the top two dipoles as was done in Figure 3. Cables L3 and L4 connect the bottom two dipoles in an identical manner. All four cables are quarter-wave (or odd multiple) cables. The output at connection points A and B is 50 ohms.

Then, these two outputs are combined in an identical manner with 75-ohm quarter-wave cables at point C. Because the output at point C is 50 ohms, standard 50-ohm cable is used between point C and the transceiver.

Many different radiation patterns can be achieved by properly connecting and phasing antennas in an array. (Note that all of the antenna arrays discussed here were shunt or parallel feed.)

While antenna design is best left to the experts, it is important to have a fundamental knowledge of the techniques employed.

Until next time—stay tuned! ■

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Demystifying RF attenuators

Creating a high-precision pad can be tricky—here's how to do it

Many different small test accessories often are needed when performing radio frequency, or RF, test and measurement work. One such item is the RF attenuator, which typically consists of several individual "pads" that can be switched in or out to increase or decrease the total attenuation. (Generally, the word "pad" refers to a single fixed attenuator.) The RF attenuator is frequently used in many test and measurement

procedures in radio communication; practical information on its construction and use is presented here.

Attenuators for audio frequencies are easily constructed out of ordinary resistors using a formula or chart to design the pad. However, RF represents a different ballgame. At high RFs, the design of the pad is much more stringent. Special resistors often are used to make a highly accurate pad. For example, a 6 dB pad might actu-

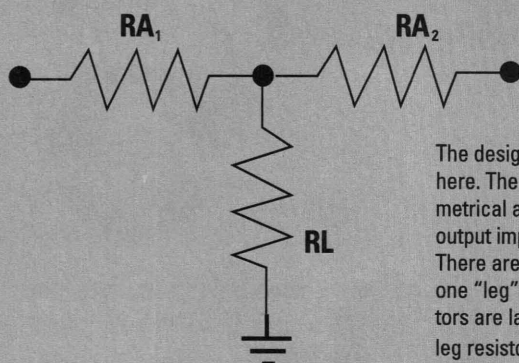
ally present 6 dB of attenuation at 1 MHz or lower but shows a significantly different attenuation at 150 MHz. If you are trying to make precise measurements, then high-precision attenuators are necessary.

Precision attenuators for RF applications are not cheap, and a variety of different attenuation levels often are needed; consequently, a well-equipped RF toolbox should contain a good variety of RF attenuators. Fixed attenuators of 3 dB, 6 dB and 10 dB are most often needed. RF attenuators are available with different RF connectors. Because of the quick connect/disconnect feature, the BNC connector is convenient for testing applications.

However, instead of BNC connectors, many RF devices feature UHF or Type N RF connectors. Unless you have the correct connector on the pad, RF adaptors will be required to make the transition from one type of connector to another. With the correct RF adaptor one can connect "anything to anything." However, the liberal use of RF adaptors can lead to problems. If possible, it is best to use an RF pad with the proper RF connector for the particular application at hand. RF pads can be connected in a piggyback fashion to provide an attenuation equal to the sum of the individual pads. However, it is best to use a single pad with the proper attenuation, if possible.

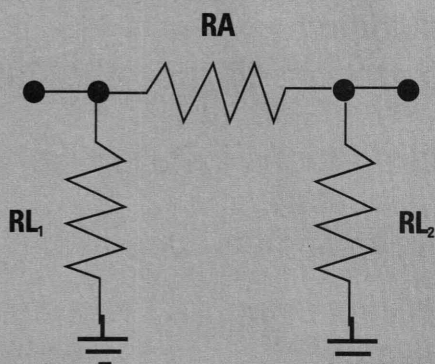
Step attenuators also are available in a variety of different attenuation levels and can be used to provide attenuation values ranging from 1 dB to 100 dB or more. Switches for step attenuators may be simple toggle switches, rocker switches, slide switches or push-button switches. By switching in the appropri-

Figure 1:



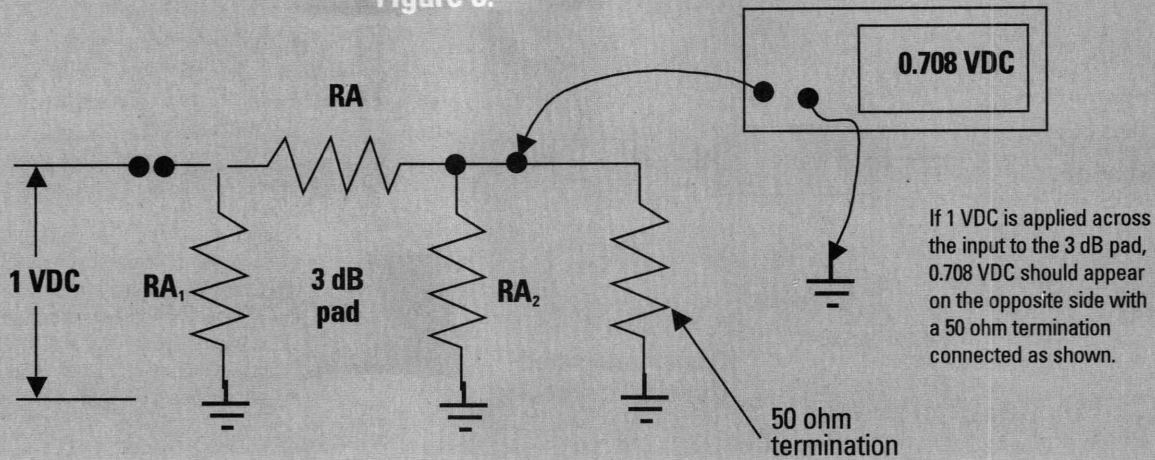
The design of the "T" pad is shown here. The input and output is symmetrical as long as the input and output impedances are equal. There are two "arm" resistors and one "leg" resistor. The arm resistors are labeled RA_1 and RA_2 . The leg resistor is labeled RL .

Figure 2:



The "pi" pad has two "leg" resistors and a single "arm" resistor. The pi pad is symmetrical as long as the input and output impedances are equal.

Figure 3:



ate attenuators, the attenuation can be increased or decreased in 1 dB steps. Attenuators may be constructed in the "T" configuration (Figure 1) or the "pi" configuration (Figure 2).

Table 1 shows the values of resistors used in the T pad for various attenuation values, while Table 2 shows the values of the resistors used in the pi pad for the same attenuation values. The attenuators are symmetrical, provided the input and output impedances are the same. In applications where attenuators are used as impedance-matching devices, they are not symmetrical.

Step attenuators are available in a variety of switching formats. They may be relay-programmable, TTL-programmable, GPIB-programmable and so on. No matter how the total attenuation is set, the principle is the same—individual pads equaling the total desired attenuation are switched into a cascade arrangement. The unused pads are simply bypassed by a straight-through connection.

Several different methods can be used to check the accuracy of an attenuator or a single pad. The simplest method is to use an ohmmeter to check the resistance. The resistance of an attenuator or pad designed for a 50 ohm system will show the same resistance on either side because they are

Table 1:			
Attenuation (dB)	RA ₁ (ohms)	RA ₂ (ohms)	RL (ohms)
3	8.5	8.5	141.9
6	16.6	16.6	66.9
10	26	26	35.1
20	40.9	40.9	10.1

This table shows the values of resistors required to make a "T" pad of various attenuation values in a 50 ohm system.

Table 2:			
Attenuation (dB)	RA ₁ (ohms)	RA ₂ (ohms)	RA
3	292.4	292.4	17.6
6	150.5	150.5	37.4
10	96.2	96.2	71.2
20	61.1	61.1	247.5

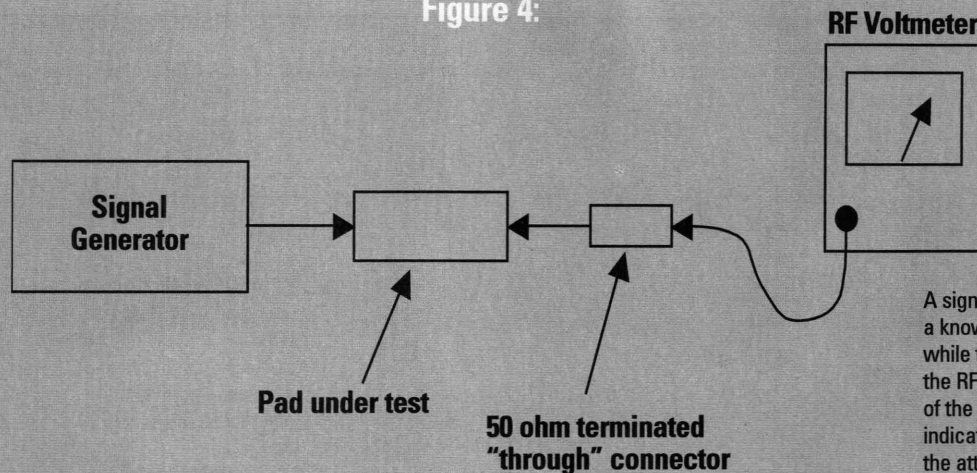
This table shows the value of the "leg" and "arm" resistors to make an attenuator for several values of attenuation in a 50 ohm system in "pi" configuration.

of a symmetrical design. This is true whether the pad or attenuator is made in the T or pi configuration. This is not true of impedance-matching pads.

To determine the resistance that should be "seen" across the input or

output of the pad or attenuator, you will need to know the resistance of the arm and leg resistors that make up the pad. The values of the arm and leg resistors for a few values of attenuation are shown in Table 1 for T pads and in

Figure 4:



A signal generator is used to feed a known signal level into the pad while the RF voltmeter measures the RF level on the opposite side of the pad. The voltmeter should indicate the generator level minus the attenuation level of the pad.

Table 2 for pi pads. Using a 3 dB pi pad as an example, Table 2 shows that the arm resistor should be 17.6 ohms, and the leg resistors should be 292.4 ohms. Thus, the resistance seen across the input or output is 150.5 ohms. If a step attenuator is composed of pi pads, the test can be conducted with the 1 dB and 2 dB pads switched in and all others switched out. Or, just the 3 dB pad can be switched in, with all others out. The ohmmeter reading should be the same no matter how the 3 dB attenuation is obtained.

Another method that can be used is

to connect a DC voltage of 1V across the input of the attenuator or pad (see Figure 3). For a 3 dB pad, the voltage appearing across the other side of the pad will be 0.708 volts. The voltage reduction factor can be determined by using the formula in Equation 1.

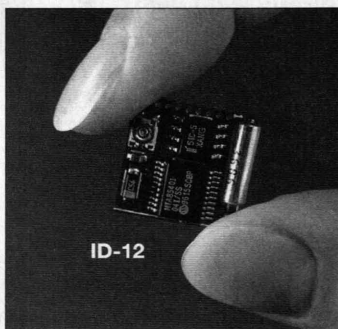
In Figure 3, 1 VDC is applied across the input to the 3 dB pad and 0.708 VDC is measured at the opposite side. It is important to note that a 50 ohm termination should be connected to the output side where the attenuated voltage is measured. If the measured voltage differs significantly

An important point to remember is to never key a transmitter into the pad. Unless the pad is designed to handle the transmitter power, it could be damaged.

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from the calculated voltage, one or more of the resistors making up the pad has changed value.

While the two previous test procedures will reveal whether the resistors comprising the pad are of the correct value, the tests do not indicate the accuracy of the pad at RF frequencies. A couple of other methods can be used for the RF test.

Figure 4 shows a signal generator connected to an attenuator on one side and a RF voltmeter connected to the opposite side through a 50 ohm terminated "through" connector. The signal

Equation 1:

$$V = 10^{\left(\frac{-dB}{20}\right)} = 10^{\left(\frac{-3}{20}\right)} = 10^{-0.15} = 0.7079$$

Where V is voltage factor and dB is pad attenuation value.

Several different methods can be used to check the accuracy of an attenuator or a single pad. The simplest method is to use an ohm meter to check the resistance.

generator is set to the desired test frequency and the output level adjusted to a convenient level, such as -20 dBm. With all of the attenuation switched out, the reference reading on the RF voltmeter indicates the insertion loss of the attenuator. This is noted as the reference mark.

Then, a level of attenuation is switched in, such as 3 dB. The RF voltmeter should drop by 3 dB. Increasing the signal generator output level by 3 dB should return the RF voltmeter reading to the reference mark obtained at the beginning of the test procedure. This test should be run at increasingly higher frequencies until a point is reached where the accuracy of the attenuator fails to meet the required accuracy. All attenuator levels should be checked at each frequency.

Another way to do the RF test is to use a spectrum analyzer that can be set for scale divisions low enough to check the lowest attenuation level to be checked. For example, to check an attenuator at 1 dB attenuation, the spectrum analyzer should be set to 0.25 dB/division. This means that an attenuation of 1 dB will change the displayed signal by 4 divisions.

Generally, the dB/division control is set to the least value that will give maximum change of the displayed signal without going off the screen. To check for 5 dB attenuation, the signal generator is adjusted to 0 dBm, and the spectrum analyzer level is set to produce a display at the top of the ref-

erence scale with the dB/division control set to 1 dB/div. Then, with the 5 dB attenuator switched in, the display should drop by 5 divisions. Increase the signal generator level by 5 dB to ensure that the display returns to the previous level. This test should be run at all the frequencies at which you plan to use the attenuator or pad.

Typical uses of pads include the "padding" of the signal generator output to make sure the signal generator always sees a 50 ohm load impedance.

When testing bandpass or reject cavities, both sides of the cavity should be connected through a pad of 3 dB to 6 dB of attenuation. Inputs and outputs of amplifiers often are padded to ensure stable operation. Pads are used in many test and measurement procedures to ensure measurement accuracy.

An important point to remember, above all, is to never key a transmitter into the pad. Unless the pad is designed to handle the transmitter power, the pad could be damaged severely, rendering it useless. Proper handling and care of the pad or attenuator will yield many years of reliable service. Handle with care! ■

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Inside line-of-sight propagation

Don't be confused: radio line-of-sight isn't always the same as optical line-of-site

When a line-of-sight path exists between two antennas, the propagation loss is equal to free-space attenuation. Free-space propagation occurs in the far-field region. Optical line of sight does not necessarily provide radio line-of-sight conditions. It depends on the clearance of the Fresnel zone. Obviously, these terms need to be defined mathematically and literally before continuing.

Free-space attenuation

Free-space propagation loss increases by 6 decibels (dB) per octave of distance from the transmitter antenna. Technically speaking, it is 6 dB per octave of distance from the point where far-field propagation begins. Free-space attenuation is found from the formula in Equation 1, where A is path attenuation in decibels, F is frequency in megahertz and D is distance in miles. It is important to note that this formula assumes that the antennas on each end are half-wave dipoles. After all, it takes an antenna to radiate a signal and one to capture it.

Figure 1 shows a free-space link between two antennas operating at a frequency of 800 MHz at a distance of 15 miles. The path attenuation between the half-wave dipole antennas is 114 dB. This is the actual attenuation between the antenna terminals.

Far-field region

The far-field region is the region in which plane-wave propagation takes place. The beginning of the far-field region can be determined from the formula in Equation 2, where F is the distance (in feet) from the antenna where the far field begins, D is the largest dimension of the antenna (in feet) and λ

is the wavelength in feet. Also, F must be much greater than both D and λ . The beginning of the far-field region should be no less than 10 wavelengths from the antenna. In the case of Figure 1, the half-wave antenna would have a dimension of 0.62 feet and λ would be 1.23 feet.

Plugging these figures into Equation 2 would yield a distance of 0.625 feet. Thus, the far-field region begins 0.625 feet from the antenna. Since F is not much greater than D or λ , the distance to the beginning of the far field should be taken as 10 wavelengths or approximately 12 feet. Relative to the 15-mile

Equation 1:

$$A = 32.7 + 20 \log F + 20 \log D$$

Equation 2:

$$F = \frac{2D^2}{\lambda}$$

Equation 3:

$$H = 2280 \sqrt{\frac{D_1 \times D_2}{F(D_1 + D_2)}}$$

Figure 1:

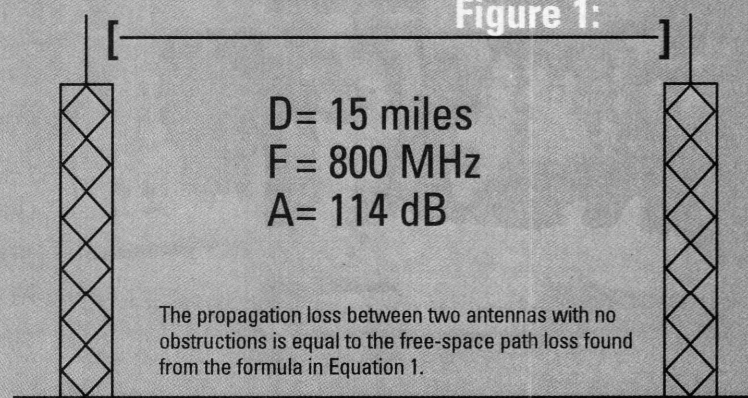
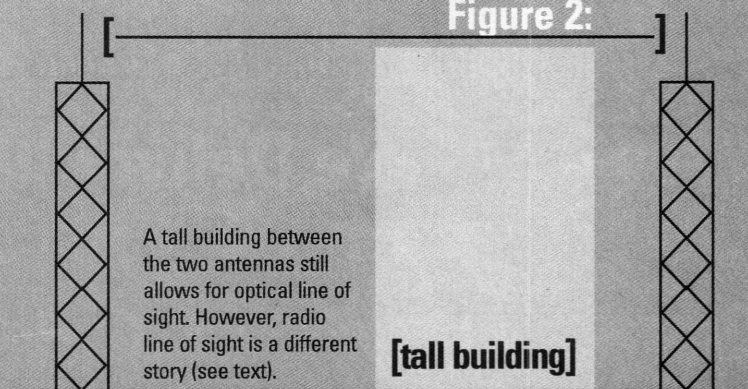


Figure 2:



distance between the two antennas, 12 feet is insignificant.

Fresnel zones

Radio line of sight and optical line of sight are different. Even over a smooth earth with no obstructions, radio line of sight is limited by the earth's curvature. Radio wave propagation extends beyond optical line of sight. So, for radio line-of-sight calculations, the earth's radius is taken to be 1.333 times or $4/3$ the normal radius. This produces a flatter earth between two fixed points.

In order to achieve true line-of-sight propagation (free space) the true line-of-sight path between two fixed antennas must clear any obstruction by more than the distance required for optical line of sight. Figure 2 shows a situation

While it isn't always necessary to achieve a line-of-sight path between two fixed points, it does greatly improve the probability of communication to do so.

where optical line of sight exists between two antennas. However, Fresnel zones exist around the radio wave and at least 60 percent of the first Fresnel zone must clear any obstruction in order to achieve minimal degradation to the radio wave for virtual free-space propagation to exist. Figure 3 shows the first Fresnel zone around the radio

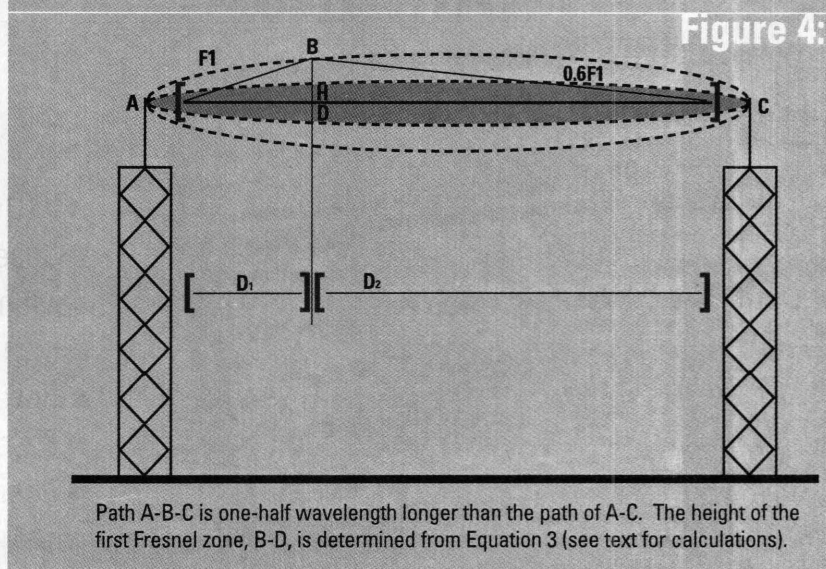
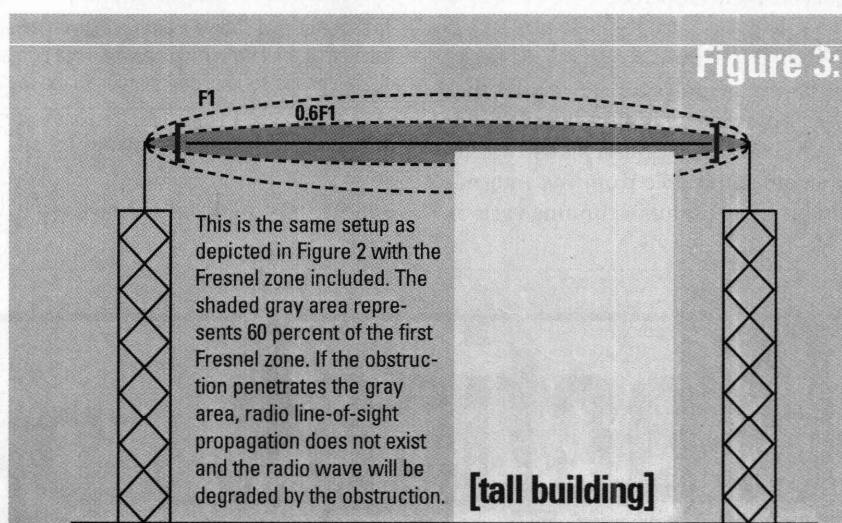
wave and 60 percent of the first Fresnel zone is shown as the gray shaded area. Notice that the tall building penetrates the Fresnel zone such that not even 60 percent of the first Fresnel zone is clear. Thus, radio line of sight does not exist and the radio wave will be degraded by the obstruction.

Equation 3 can be used to calculate the Fresnel zone around the radio wave at any given point between the receiver and transmitter antennas. In Equation 3, H is the height (in feet) of the first Fresnel zone above optical line of sight. D_1 is equal to or less than D_2 and is measured in miles. F is the

frequency in megahertz.

Figure 4 shows the first Fresnel zone is an ellipsoid surrounding the straight line between the two antennas. The sum of the distances between the two antennas and a common point on the ellipsoid is one-half wavelength greater than the straight line between the two antennas. For example, in Figure 4 the length of A-B-C is one-half wavelength greater than length A-C.

Suppose that the straight-line distance (A-C in Figure 4) between the two antennas is 25 miles, D_1 is 10 miles, and D_2 is 15 miles, while the operating frequency is 800 MHz. Plugging these

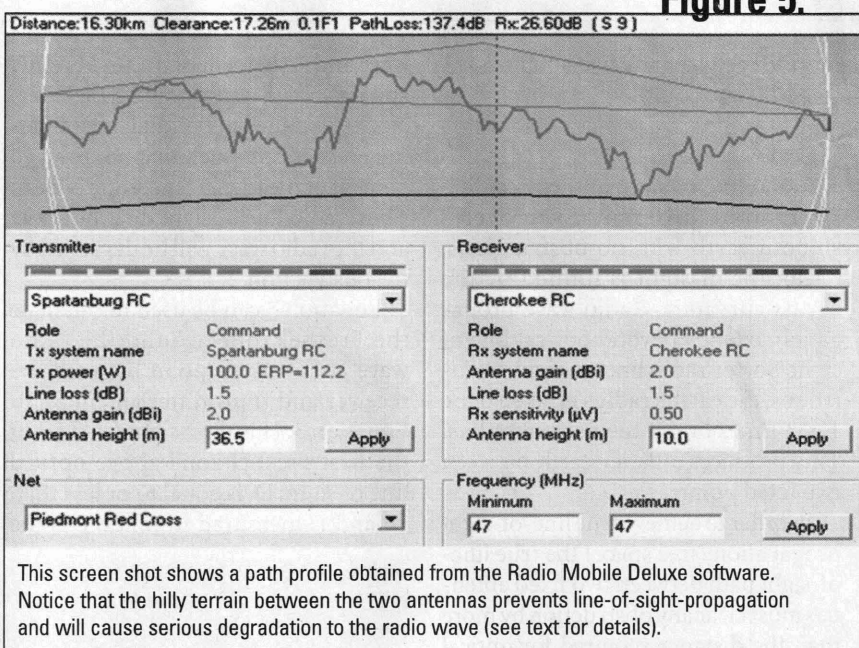


numbers into Equation 3 yields a solution of 197.5 feet for the first Fresnel zone at distance D. The wavelength at 800 MHz is 1.2303 feet. The straight-line distance (A-C) is equal to 25 miles (132,000 feet or 107,290.9 wavelengths).

Using the Pythagorean theorem and substituting 197.5 feet for B-D, we find that the length of A-B is 52,800.37 feet or 42,916.66 wavelengths. Similarly, length B-C is calculated to be 79,200.25 feet or 64,374.74 wavelengths. The sum of A-B and B-C is 107,291.4 wavelengths, 0.5 wavelengths longer than A-C. This simply proves that the calculation for the height (B-D) of the first Fresnel zone is correct.

Path profile

Figure 5 shows a path profile that was obtained using software called Radio Mobile Deluxe. This software is downloadable free from the Internet and is quite useful in running various

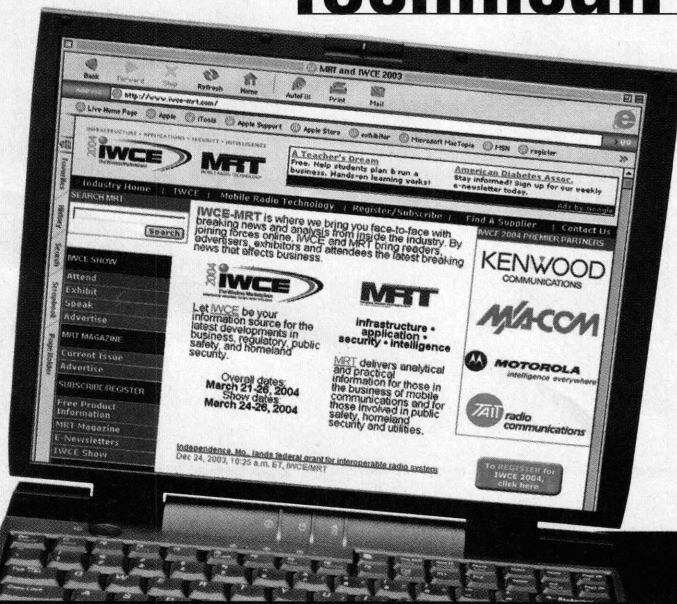


communication scenarios using actual terrain data.

Figure 6 shows the boundary of 60

percent of the first Fresnel zone. Notice that 60 percent of the first Fresnel zone would not even clear smooth earth.

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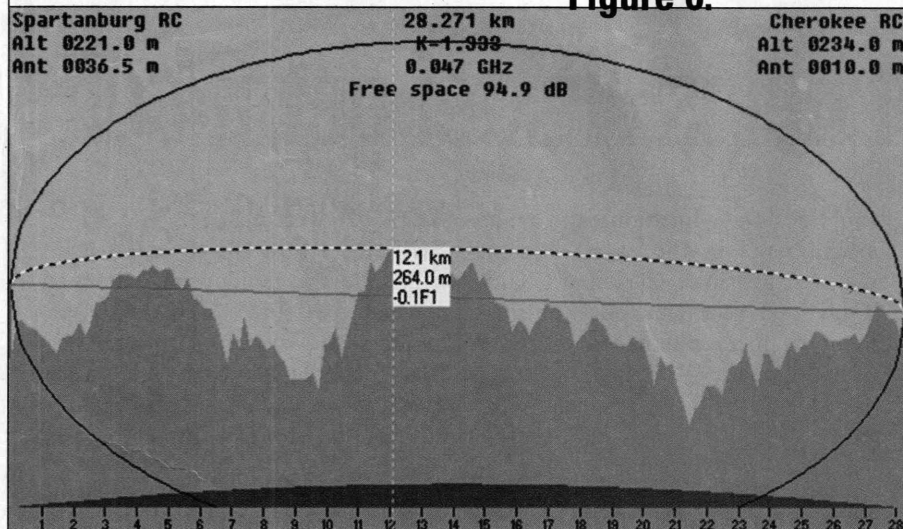


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Figure 6:



This screen print shows the curvature of the earth at 1.333 times the normal radius. The terrain profile is clearly shown along with the boundary of 60 percent of the first Fresnel zone. This was obtained with the Radio Mobile Deluxe freeware.

Smooth earth is indicated by the black curvature at the bottom of the figure. With the hilly terrain between the two antennas, the signal would encounter severe attenuation.

Summary

As the frequency of operation increases, the boundary of the Fresnel zone draws in closer to the straight line. That is, at very high frequencies the propagation approaches optical line of sight. While it isn't always necessary to achieve a line-of-sight path between two fixed points, it does greatly improve the probability of communication to do so.

This is especially important in mi-

crowave links in order to achieve maximum reliability. Using a program such as Radio Mobile Deluxe with the proper terrain data (also free from the Internet) will allow one to determine necessary antenna heights on each end as well as other pertinent information for establishing a reliable communication link.

Until next time—stay tuned!

¹ Radio Mobile Deluxe was written by Roger Coude of Canada and made available as freeware.

Editor's note: The web site for Radio Mobile Deluxe is www.cplus.org/rmw/english1.html. ■

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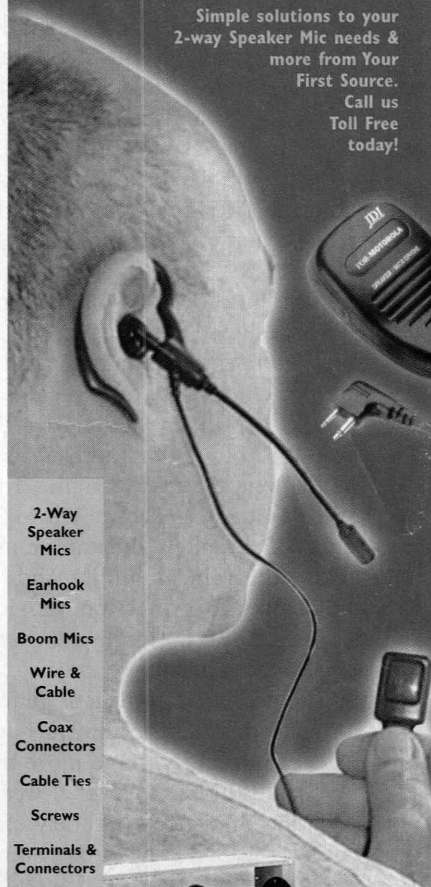


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Mayday! 'Mayday' 'Mayday!' Mayday! Mayday! Mayday!

This call went out for firefighters to evacuate the World Trade Center before the first tower collapsed, but many died because they allegedly never heard it. Even if a court never considers a lawsuit that blames the radio system, the dynamics of the mobile industry will change forever.

by Donny Jackson

More than 300 New York City firefighters lost their lives in the Sept. 11, 2001, attack on the World Trade Center. Families of 12 of those firefighters filed a \$5 billion lawsuit alleging their relatives died because radios failed to receive evacuation orders before the towers collapsed.

Although it is very possible a jury may never hear the case, the lawsuit's underlying message of increased accountability is being heard clearly throughout the mobile radio industry.

"With the emergence of the first-responder industry ... people are going to be less tolerant of equipment that is not effective for use," said Richard Salem, lead attorney for the families in the lawsuit against the City of New York and Motorola, the parties the plaintiffs claim were responsible for ensuring that the radio system worked on Sept. 11, 2001 (see sidebar). "We don't even like a lost call when we use a cell phone. And there's no reason for it—the fun-

damental technology and equipment are there."

Given this, Salem said it is "inevitable" that there will be more lawsuits filed when public-safety officers are injured or killed in future incidents when radio systems do not work properly—especially after large damages are awarded in such a case.

"Once firefighters in Peoria, Santa Monica and Austin realize that this happened in New York, I think they will start asking questions," he said.

And the possibility of large awards is very real, because public-safety radio communications can have a life-or-death impact on those willing to risk their lives for the very jurors who would assess damages, according to Karl Protil, a shareholder in the Rockville, Md., law firm of Shulman, Rogers, Gandal, Pordy & Ecker.

"Human nature being what it is, if you have a poor widow or children without a father, it's going to impact a jury ... even though the judge will instruct the jury not to let it prejudice

them," Protil said. "It has to be a concern for anyone trying to defend one of these cases. Anyone who says otherwise is lying, I think."

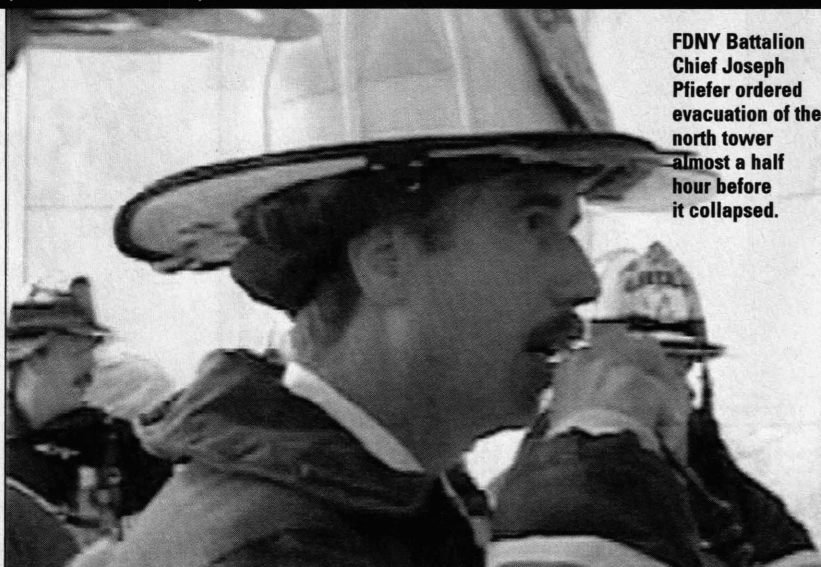
The specter of legal action could permeate almost every step in the process of providing radio support for public-safety personnel, from contractual practices to day-to-day operations to a willingness to adopt innovations. And many of these potential changes have unwanted implications, according to Vincent Stile, chairman of the Association of Public-Safety Communications Officials (APCO) International.

"If we always have to be concerned about the possibility of a lawsuit, ... that would put a damper on everything that we do," he said.

The process begins with the awarding and wording of contracts. It's generally accepted that no radio system can provide 100% coverage of any significant geographic area all the time, even with unlimited funds to spend. The chal-

9/11 Timeline Title Here

(all times are EDT)



FDNY Battalion Chief Joseph Pfiefer ordered evacuation of the north tower almost a half hour before it collapsed.

8:45 a.m.: A hijacked passenger jet, American Airlines Flight 11 out of Boston, Massachusetts, crashes into the north tower of the World Trade Center.

9:03 a.m.: A second hijacked airliner, United Airlines Flight 175 from Boston, crashes into the south tower of the World Trade Center.

9:32 a.m.: FDNY Assistant Fire Chief Joseph Callan issues evacuation order for all firefighters in the north tower. The "overwhelming majority" of the firefighters do not hear the order, according to the lawsuit.

9:59 a.m.: The south tower of the World Trade Center collapses.

10:00 a.m.: After the collapse of the south tower, FDNY Battalion Chief Joseph Pfiefer issues emergency evacuation order for the north tower. According to the lawsuit, most firefighters were unable to hear this call.

10:29 a.m.: The World Trade Center's north tower collapses.

Source: CNN.com, Lawsuit against the City of New York and Motorola

"If you have a poor widow or children without a father, it's going to impact the jury ... It has to be a concern for anyone trying to defend one of these cases."

—Karl Protil, attorney

lenge is much greater for government entities faced with the prospect of meeting the communications needs of first responders within budgets already stretched by unfunded homeland-security mandates from the federal government.

"From an economic model, the question becomes what is plausible to spend," Salem said. "There should be an economical way to supply emergency-service personnel with a communication system that will work."

In light of the New York City lawsuit, ensuring that systems are designed correctly could lead to the hiring of more independent consultants to help with this phase of communications projects, according to Dana Hansen, superintendent of communications for the Denver police department. And virtually everyone interviewed for this article agreed that future specifications and contracts likely will be more carefully worded, while vendors will take additional measures to make sure the capabilities of the system being installed are clearly understood.

"I also think it will spur the manufacturers to ensure that the product will be exactly what they represented," Protil said.

Exactly how much coverage is

Lawsuit may not reach court, but allegations persist

Were the deaths of firefighters in the collapse of World Trade Center simply unfortunate byproducts of a terrorist act, or could they have been avoided if the Fire Department of New York had a better radio system in place on Sept. 11, 2001?

This question is at the heart of a \$5 billion lawsuit against New York City and Motorola that likely will never be heard by a jury. All 12 plaintiff families of deceased firefighters have sought aid from a special fund for Sept. 11 survivors, which may make them ineligible to pursue such litigation.

If the case isn't heard, it would be unfair, according to Richard Salem, lead attorney for the plaintiffs in the case.

"We realize that the court and the public, in general, would prefer that the travesty and tragedy of 9/11 be put into the annals of history," Salem said. "However, did Congress intend to exonerate wrongdoers for their actions, be it before 9/11 or during 9/11 [by creating the victims' fund]? I don't think so."

And Salem said there was wrongdoing, which is why plaintiffs believe they have grounds for a lawsuit. Normally, litigation is not allowed in firefighter fatalities, because the job is recognized as inherently dangerous. However, litigation is allowed when the deaths are caused by the city knowingly giving ineffective equipment to firefighters or because of a legal violation. Both of these conditions apply in this case, Salem said.

Problems with the radio system were well known. After the 1993 bombing of the World Trade Center, FDNY's inability to communicate with firefighters on the upper floors of the building was cited as an issue that needed to be addressed. However, essentially the same system was in place on Sept. 11, 2001, according to Salem.

"Both Motorola and FDNY knew that the Saber radios in the hands of firefighters on 9/11 did not, and could not, work in high-rise buildings," Salem said. "Then, they sent them into high-rise buildings and told them to go up, up, up."

"The law requires an employer to provide equipment that is free from recognized risk and hazard and suitable for safety and use within its intended environment. To the

extent that an employer can replace gear [to address a problem], it is required to do so."

In fact, FDNY tried to replace its radios after the WTC bombing, spending at least \$13.9 million for 3818 XTS 3500 radios, which Motorola promoted as feature-rich digital handsets that could be substituted for the discontinued Saber line of products. The XTS 3500 radios finally were deployed two years after the agreement was signed, only to be shelved after a week of complaints, including alleged nontransmission of a firefighter's call while trapped in the basement of a house in Queens in March 2001.

After reviewing the XTS 3500 contract, then-New York City Comptroller Alan Hevesi said the absence of competitive bids for a new radio system broke the city's contracting rules—a violation that provides further grounds for plaintiffs to pursue legal action, according to Salem.

Hevesi also claimed the XTS 3500 "may

"How could you design a system to handle that?"

—John McFadden, Motorola VP, on 9/11 circumstances

have [been] a prototype, rather than a fully developed radio" and that it was "irresponsible and dangerous" to deploy the radios before they were tested in the field.

"It appears that Motorola allowed the FDNY to unknowingly become a large-scale testing ground for the XTS 3500 radio to the possible detriment of New York City firefighters," Hevesi wrote in a May 2001 letter.

Motorola asserts it properly followed procurement procedures and conducted all tests requested by FDNY, according to spokeswoman Pat Sturmon.

"Before the radios were ever deployed, they went through an extensive testing pro-

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More than 300 firefighters died in the World Trade Center attack, many of them needlessly, says a lawsuit filed against New York City and Motorola.



"It's premature to talk about it now, but if everybody's worried about getting sued, nobody will do anything. And then, the public really loses."

—Vincent Stile, APCO chairman

enough remains a question without a clear legal answer, and each hole in radio coverage represents a potential source of litigation. But Stile said he believes there should be no obligations for governments to accept the most complete system if the price is out of line, noting that "sometimes, the Cadillac [system] has bells and whistles that you don't need."

Perhaps a more sensitive question is whether an entity is legally liable if it decides not to appropriate available funding to resolve a known problem in the radio system. In the interim, there are instances when communications leaders have little choice but to hold their breath when a potential communications problem becomes evident, Hansen said.

"We've experienced that in Denver with some of the new buildings being built," she said. "I just try to put people

on notice and keep the issue alive, and pray that nothing bad happens until we can get a grant to fix it."

In other cases, in-house communications personnel are able to provide temporary fixes to dead spots in radio coverage by deploying portable repeaters and other "resourceful" techniques, Stile said. Indeed, one of the strengths of many public-safety communications departments is the ability to "improvise" solutions when there is a shortage of money or time to address a problem, he said.

But such actions may not be appreciated as much in the future, as they could create legal issues involving vendors' warranties, Prottil said. While a company may be willing to meet performance standards for its equipment, it may not be willing to accept respon-

ity if the system is changed by the owner.

I think that's what's going to happen, as well as a re-evaluation of the [vendor-customer] relationship," Stille said. "The practice of patchwork solutions is going to have to stop."

Another possibility is that vendors will seek greater input into day-to-day operations and maintenance of the radio systems. Hansen said she hasn't heard of such situations yet, but she can imagine a time when vendors will want to ensure that software releases are installed in a timely manner and that emergency personnel are trained properly to use the equipment.

"A lot of time, it comes down to raining, from dispatchers to ... people in the field," Hansen said. "Many times, it's not a technical problem, it's a people problem."

In other instances, problems are more elusive and attributable to Mother Nature. While a factor as stagnant as geographic terrain can greatly impact the ability to receive radio signals, it also can be altered by something as dynamic as the growth of leaves on trees. Indeed, Stille said he annually has to adjust the power of radio signals in part of his Suffolk County, N.Y., communications system to offset the fact that leaves growing on trees in the summer act as mini-antennas and absorb a noticeable portion of the signals.

In Stille's case, his considerable experience has allowed him to learn about this situation and its impact on the public-safety radio system. However, Hansen questions whether it will be fair legally to require all communications leaders to be able to deal with such issues.

"At that point, you have to ask, 'Is it reasonable?'" Hansen said. "Is it reasonable to expect someone new to that? I don't think so."

Whether a jury agrees is anybody's guess. And such uncertainty has the potential to shake the public-safety organiza-

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cedure, including water submergibility," Sturmon said. "And Motorola did meet FDNY's very stringent specifications."

In terms of the events of Sept. 11, Motorola initially provided the following prepared statement from John McFadden, vice president of sales for the North America Group's northern division.

"The terrorist attacks on the World Trade Center were horrible and tragic, and our hearts go out to those who lost family members and loved ones," McFadden said.

"However, there is no basis for the claims against Motorola, and we will defend ourselves against these false accusations."

"As to the events on Sept. 11, there are personal and news-media reports that demonstrate that New York City firefighters were able to communicate through Motorola two-way radios. In fact, lives were saved that day because of these communications."

In a subsequent interview, McFadden said the primary change in FDNY's radio system after the 1993 World Trade Center bombing was that the Port Authority of New York and New Jersey installed a repeater on a nearby building to enhance radio coverage on the towers' upper floors. But "no one knows what happened to that repeater" on Sept. 11, McFadden said.

But it would not have mattered whether the repeater worked or the XTS 3500s had been used on Sept. 11, McFadden said. The Saber radios worked as designed, as reflected by communications with a firefighter near where the plane hit the south tower.

The problem is that New York City has just one channel for fireground communications, which is adequate to handle the peer-to-peer communications necessary for most incidents, McFadden said. However, on Sept. 11, hundreds of firefighters tried to talk on the one channel, which meant most communications could not be transmitted, he said.

"How could you design a system to handle that?" McFadden said.

But police-department radios worked, proving the day's unique circumstances could be overcome, Salem said.

"Either there is technology that works in high-rise buildings or there isn't. If there is, then it needs to be in there. If there isn't, at the very least, we need to warn the firefighters ... before they risk their lives." ■

Road to NYC radio litigation

February 1993—Bombing at World Trade Center reveals a number of problems in the communications system used by FDNY. A report cites the need to address the inability to communicate with firefighters in the upper floors of the World Trade Center using Motorola's Saber I analog "handi-talkies" and in-car repeater systems.

April 1997—City of New York agrees to purchase 750 Saber I radios from Motorola for \$2.88 million over three years in a sole-source replacement contract.

May 1998—Motorola informs New York City that the Saber I radios will be discontinued. The vendor offers to substitute the Saber radios with XTS 3000 models. During the same year, Motorola introduces the Saber III product.

February 1999—Motorola says it no longer manufactures Saber III radios and offers to substitute them with XTS 3500 radios, which are new digital models.

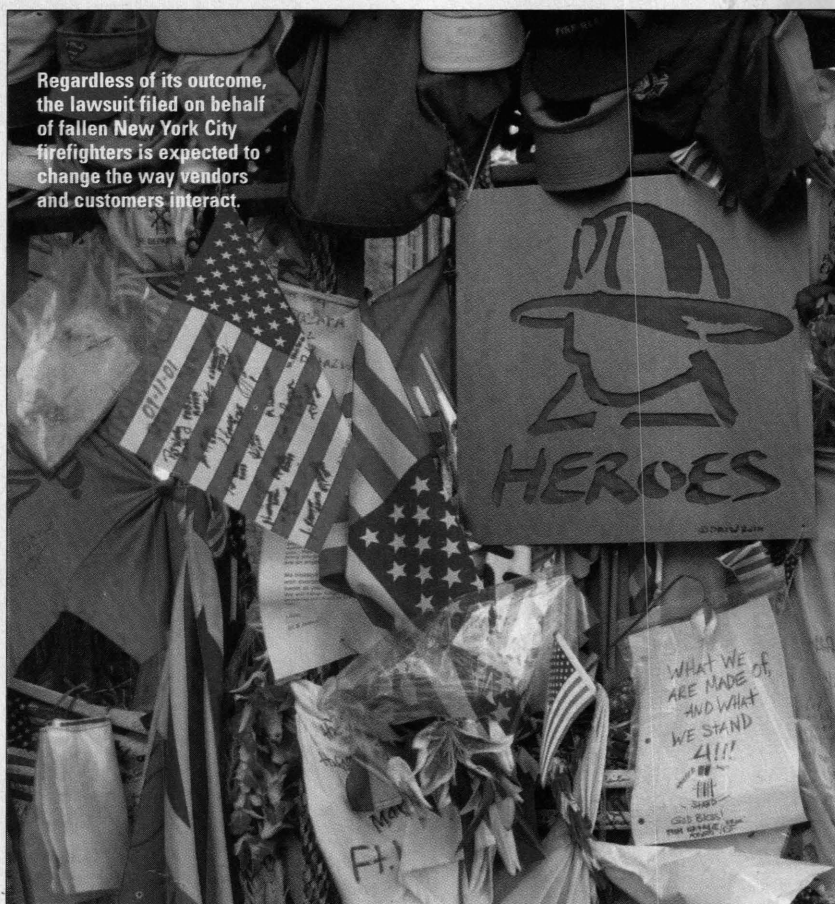
December 1999—Over a three-month period, FDNY places \$13.9 million in orders with Motorola for 3,818 XTS 3500 radios as a substitute for the \$2.88 million contract for 750 Saber I radios in 1997, meaning no competitive bids are sought. First 2,700 radios delivered are returned because they were not waterproof.

March 2001—FDNY puts XTS 3500 radios into service. After numerous complaints and a life-threatening incident involving a firefighter in Queens, the radios are pulled from service a week later.

April 2001—New York City Comptroller Alan Hevesi releases a review that states the XTS radios were not field tested before being put into service in March 2001 and orders payments to Motorola be stopped. Hevesi also claims FDNY broke contracting rules by purchasing a new radio system without competitive bids as a substitute for a contract for replacement radios. A week later, a Motorola official testifies there is nothing technically wrong with the radios.

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Regardless of its outcome, the lawsuit filed on behalf of fallen New York City firefighters is expected to change the way vendors and customers interact.



"[Potential litigation] makes it compelling for public-safety manufacturers to rise to a higher level of quality. And municipal, state and federal government officials have to demand that."

—Richard Salem,
lead attorney for 9/11 plaintiffs

tions to adopt only proven systems while eschewing innovative technologies. Hansen said she does not believe this will occur but cautions that backup plans need to be in place in case the new technology proves to have bugs that prevent it from being deployed.

"I think you've got to have a fallback when you're beta testing until you know the new system works," she said.

Stile said he hopes the industry never adopts a litigation-avoidance mindset that retards innovation in mobile-radio communications, because that eventually could jeopardize the lives of emergency responders and the civilians they are trying to protect.

"I would hope we don't go backward; that would be the wrong thing to do," Stile said. "It's premature to talk about right now, but if everybody's worried about getting sued, nobody will do anything. And then, the public really loses."

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May 2001—Hevesi asks Motorola to return the 3,818 XTS 3500 radios sold to the City of New York. Hevesi claims the XTS 3500 radios were an undeveloped prototype when sold to the City in 1999 and that Motorola refuses to identify other buyers of the radios.

September 2001—Using essentially the same Saber radio technology that was used during the 1993 bombings, firefighters allegedly are unable to hear evacuation orders before the north tower of the World Trade Center collapses.



Hevesi

Sources: New York City Comptroller's Office, Lawsuit against City of New York and Motorola

Regardless of whether lawsuits generate large damage awards in the near future, everyone interviewed agreed that the potential of litigation raises the stakes of the already-serious business of public-safety radio communications to another notch for everyone involved—vendors, elected officials, communications staffs and responders who need to be trained.

"It makes it compelling for public-safety manufacturers to rise to a higher level of quality," Salem said. "At the municipal, state and federal government levels, officials have to demand that."

It's an evolution that has been in the making for some time, particularly in the wake of the events of Sept. 11, Hansen said.

"I think people are going to pay more attention more—from all sides. We're demanding more from our vendors ... but it's important for us to get up my end of the bargain, and